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A MAGNETIC WHEEL STORE FOR RECORDING TELEVISION SIGNALS

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The magnetic recording of signals on the periphery of a wheel or drum, coated with a magnetic material, has been known for some considerable time: it is used, for example, to delay acoustic signals for producing reverberation effects (ambiophony), and for information storage in certain electronic computers. The article below describes a system by which television signals can be magnetically recorded on the rim of a wheel, and subsequently displayed as and when required. The system is likely to find an important application in radiology.

Video tape recording

The first point to be noted about the magnetic recording of television picture signals is that it involves frequencies very much higher than are encountered in sound recording. In the latter the top frequency is about 20 kc/s, whereas for television pictures (with 625 lines) it is necessary to go up to about 5000 kc/s.

A sinusoidal electrical signal is impressed on to the magnetic tape with a particular wavelength. Between this wavelength λ, the frequency f and the writing speed v_s (the speed of the tape relative to the recording head) the following relation exists:

$$v_s = \lambda f.$$

In order to keep down the writing speed the minimum wavelength λ_{min} must be chosen as short as possible. A lower limit is set to λ_{min} by the reproduction quality, which declines rapidly when λ_{min} drops below 5 μ. At $\lambda_{min} = 5\ \mu$ and $f = 5$ Mc/s the writing speed v_s is 25 metres per second. If the recording head is stationary, as it always is in sound recording, the writing speed is the linear speed of the tape past the head. There are obvious objections to a tape speed of 25 m/sec. One method of video tape recording ¹) gets around this difficulty by making the magnetic head rotate, thus allowing the tape speed to be reduced considerably below the writing speed.

Recording on a magnetic wheel

Where the object is not to record a whole series of television pictures (part of a television programme, for example) but only to "store" one or a few television frames in a memory device, the latter can be given a form that readily allows a high writing speed to be used. We refer to the form of a wheel, analogous to the acoustic delay wheel used to produce ambiophonic and other sound effects ²). The periphery of the wheel is provided with a coating of magnetic material, and at a very short distance from it a recording or "writing" head is mounted which, for video purposes, also serves as the playback or "reading" head. Such a wheel, which we shall presently describe *in extenso*, can easily be given a peripheral speed of some scores of metres per second. This ensures a sufficiently faithful reproduction of the *high* frequencies.

The trouble here arises at the other end of the spectrum, at the *low* frequencies. For example, at $f = 100$ c/s and $v_s = 25$ m/sec, the wavelength is 25 cm. This is much longer than the length l of the head, which is about 1 cm (*fig. 1*). The result is that the magnetic flux of the magnetized layer — in so far as it corresponds to the low frequencies — no

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¹) C. P. Ginsburg, Comprehensive description of the Ampex video tape recorder, J. Soc. Mot. Pict. Telev. Engrs. **66**, 177-182, 1957.

²) Philips tech. Rev. **17**, 259, 1955/56, and **20**, 325, 1958/59.

longer passes through the head but around it. Thus, as the signal frequency decreases, the output signal not only shows the familiar gradual drop of 6 dB per octave (because with increasing wavelength the contained magnetic flux varies more slowly with time) but also, when λ is several times longer than l, it begins to fall very rapidly to zero.

The difficulty, then, is that in video recording the ratio of the highest to the lowest signal frequency is particularly large, being about 10^5 — against 10^3 in the case of sound. The difficulty can be circumvented, however, by recording instead of the video signal itself a "carrier wave" modulated by the video signal.

Before describing this system we should point out that direct (i.e. unmodulated) recording on a magnetic wheel does produce useful results for certain purposes. Visitors to the 1958 Photokina Exhibition at Cologne who accepted the invitation to be photographed in the Philips stand, saw themselves appear, at the moment the shot was taken and for some time after, on two television screens. A magnetic wheel store played an essential part in this stunt, as explained in the caption to *fig. 2* [3]).

In 1959 the magnetic wheel was demonstrated at the 9th Radiological Congress in Munich (see end of this article) and at the Radio and Television Show in Brussels. At the latter the visitors saw a moving picture of themselves on one television screen and a stationary picture on another. The first set was connected directly to a television camera, the second via a magnetic wheel.

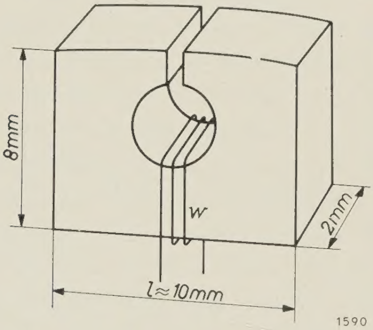

Fig. 1. Form and general dimensions of a write-read head, with schematically-represented winding *w*. The core is of ferroxcube IV. The magnetic layer travels in the direction of the length dimension *l*.

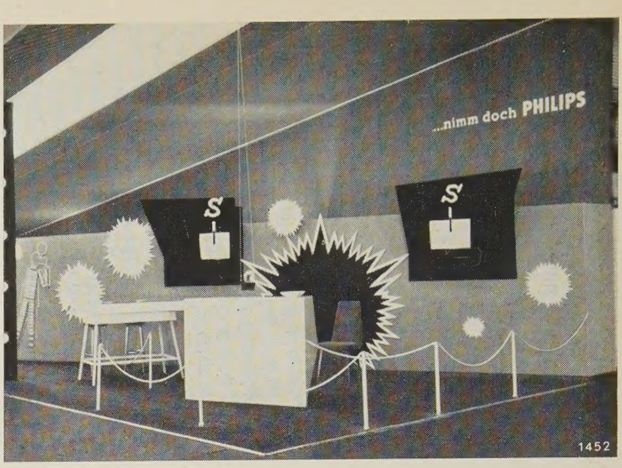

Fig. 2. Philips stand at the 1958 Photokina Exhibition in Cologne. Members of the public were invited into the stand to have a flash-photograph taken of themselves, the photograph to be posted on to them later. Immediately after the flash, a picture of the subject appeared simultaneously on two television screens *S*, and remained there for some time after the subject had left the stand.

How this television picture was produced was not disclosed, and must have mystified many. It was done as follows. When the subject took his seat in front of the photographer's camera, he was at the same time within the field of view of a concealed television camera. The light from the flash bulb made a phototransistor conductive, which in turn actuated a relay, leading successively to the erasure of the picture already stored on a concealed magnetic wheel, and to the recording and display of the new picture [3]).

Frequency modulation system

To avoid the difficulties involved in recording a signal having a frequency ratio of $10^5 : 1$, the signal can be made to modulate a carrier wave. As we shall see, the spectrum of the modulated carrier then has a much smaller frequency ratio.

Frequency modulation is the most appropriate system for the purpose [4]). In contrast to amplitude modulation, it is possible in this system to suppress modulation noise to a great extent, and moreover the noise present is so distributed over the spectrum as to cause much less interference in a television picture. It will be useful to deal at greater length with these two reasons for preferring frequency modulation.

The output signal of a magnetic recording is always modulated in amplitude by noise. This modulation noise is partly due to the fact that the distribution of the grains of the magnetic coating is not perfectly uniform [5]). Other causes are variations in the thickness of the coating, dust particles between the coating and the head, and — in the present case— imperfect roundness of the wheel. All these imperfections, then, give rise to undesired amplitude modulation. If the video signal were also present as amplitude modulation on the carrier, it would not be possible to separate these two modulations from one another. When frequency modulation is used, however, nearly all the amplitude modulation can be removed by means of a limiter, leaving an almost purely frequency-modulated signal and eliminating the above-mentioned noise contributions.

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[3]) The credit for this idea, and for part of its technical realization, goes to J. F. van Oort of the Philips Exhibition Department.

[4]) C. E. Anderson, The modulation system of the Ampex video tape recorder, J. Soc. Mot. Pict. Telev. Engrs. **66**, 182-184, 1957.

[5]) See e.g. D. A. Snel, Magnetic sound recording, Philips Technical Library 1959.

The second advantage of frequency modulation again relates to noise. In *fig. 3a* the centre frequency of the frequency-modulated carrier is denoted by f_c, whilst Δf denotes a small band of the noise spectrum about an arbitrary frequency f_n. After detection, this band comes within the video spectrum, as shown in fig. 3*b*, i.e. around the video frequency f_c—f_n. Now, the noise power ΔP_n in this band is proportional to $(f_c - f_n)^2$ (in contrast to amplitude modulation, where ΔP_n is independent of f_c—f_n). This proportionality is a favourable circumstance, experiments having shown that noise in a television picture is more troublesome the lower are the frequencies of the noise components for the same ΔP_n [6]).

The frequency-modulation system should be arranged such that the ratio of the highest to the lowest frequency in the signal to be recorded is very much smaller than in the video signal, without the highest frequency appreciably exceeding the highest video frequency. These requirements are fulfilled if the instantaneous frequency of the recorded signal is, say, 7 Mc/s for the brightest white in the picture, 5.5 Mc/s for black, and 5 Mc/s for the peak of the synchronizing signals ("blacker than black"); see *fig. 4*.

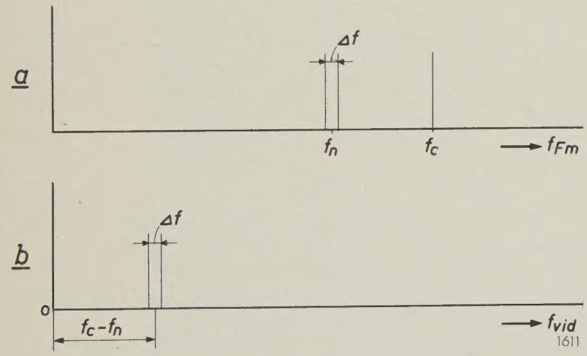

Fig. 3. *a*) Of the noise spectrum of a frequency-modulated signal, with centre frequency f_c, the diagram shows a narrow band Δf about an arbitrary frequency f_n. After detection, this band appears in the video spectrum as illustrated in (*b*). The noise contribution of Δf is proportional to $(f_c$–$f_n)^2$, and is hence smaller for low video frequencies than for high. As a consequence, the noise in the picture is less troublesome.

What the frequency spectrum of a sinusoidally frequency-modulated signal will look like depends to a large extent, of course, on the modulation index *m*, defined as the ratio of the frequency deviation Δf to the modulation frequency [7]). As a rough approximation, let us assume that the video signal is sinusoidal; given 5 and 7 Mc/s as the extreme values of the instantaneous frequency the centre frequency will then be 6 Mc/s and the frequency deviation 1 Mc/s. If *m* is greater than about 25 (i.e. in our case $f_{\rm vid} < 40$ kc/s) the spectrum will then consist mainly of numerous weak lines within the extremes 5 and 7 Mc/s (*fig. 5a*). For $m = 3$ ($f_{\rm vid} = \frac{1}{3}$ Mc/s) the lines are much farther apart, but those beyond 5 and 7 Mc/s are of little consequence (fig. 5*b*). At $m = 0.2$, however, corresponding to the highest video frequency $f_{\rm vid} = 5$ Mc/s, the spectrum is made up virtually of only three lines, at 6, 6 + 5 and 6 — 5 Mc/s (fig. 5*c*), two of which are thus far outside the 5 and 7 Mc/s limits.

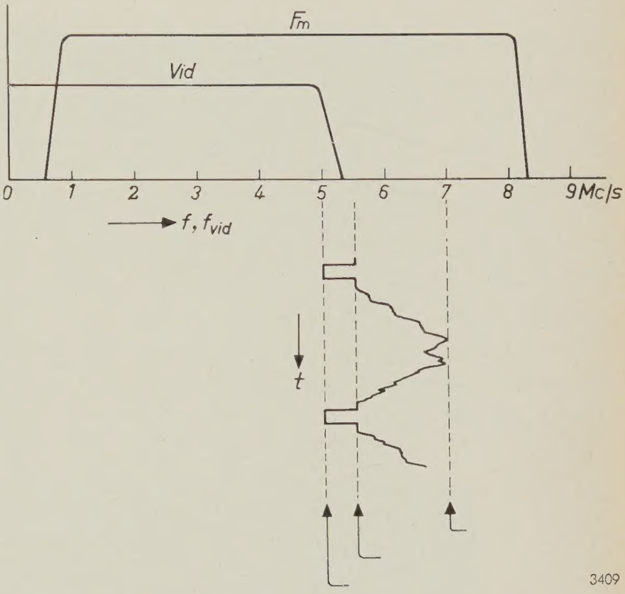

Fig. 4. *Vid* frequency range of the video signal for a television picture with 625 lines. *Fm* frequency range of the signal to be recorded, whose instantaneous frequency is 7 Mc/s in the brightest white, 5.5 Mc/s in the black, and 5 Mc/s in the "blacker-than-black" (peak of sync signals).

The latter does not mean that, for recording, we must reckon with 6 + 5 = 11 Mc/s as the highest frequency. It is sufficient if we take the lower sideband plus part of the upper sideband, up to about 8 Mc/s (curve *Fm* in fig. 4). The lower limit can

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[6]) Amongst the extensive literature on the noise nuisance in television pictures, mention may be made of: L. Goussot, Le brouillage des images de télévision par les signaux parasites, Onde électr. **39**, 352-361 and 690-700, 1959 (No. 386 and No. 388/389). This also quotes references to other articles on the subject.

[7]) See e.g. Th. J. Weijers, Frequency modulation, Philips tech. Rev. **8**, 42-50, 1946, in particular fig. 4.

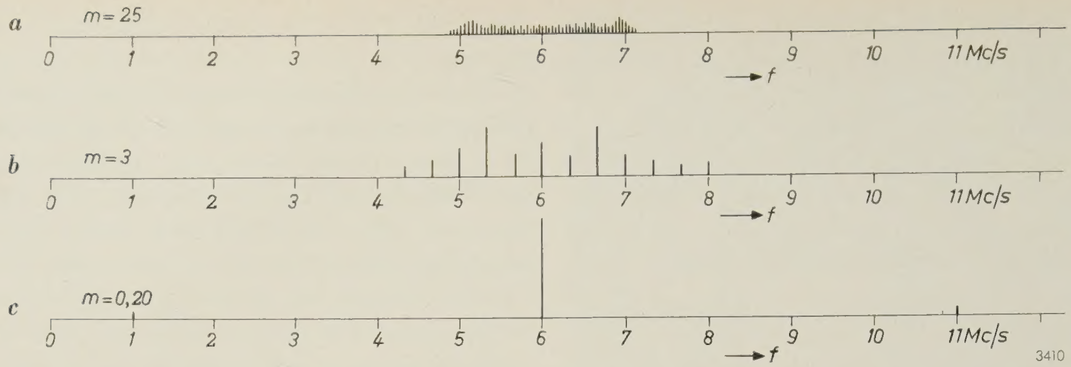

Fig. 5. Frequency spectra of a sinusoidally frequency-modulated signal, with centre frequency 6 Mc/s, frequency deviation 1 Mc/s and modulation index m = 25, 3 and 0.20, respectively.

usefully be chosen between 0.5 and 1 Mc/s. The frequency ratio of the modulated signal is then of the order of 10 : 1, instead of the ratio of 10^5 : 1 for the video signal. Expressed in octaves, the frequency range to be recorded is thus reduced from 17 to 3 or 4 octaves. True, the highest frequency for recording is now 8 Mc/s instead of 5 Mc/s, but this presents no major difficulties for recording on a wheel store.

The quality at present achieved with the aid of frequency modulation appears from *fig. 6*, which shows a photograph of a resolution chart obtained on a monitor via a magnetic wheel. Further improvements in quality may be expected in the near future.

Mechanical features of the magnetic wheel

To keep the wheel store as small as possible, it was decided in the design stage to record one television frame on the periphery; one frame lasts $^1/_{50}$ second, hence the wheel must turn at 3000 revolutions per minute.

As we have seen, the highest frequency in the signal to be recorded is 8 Mc/s. The shortest useful

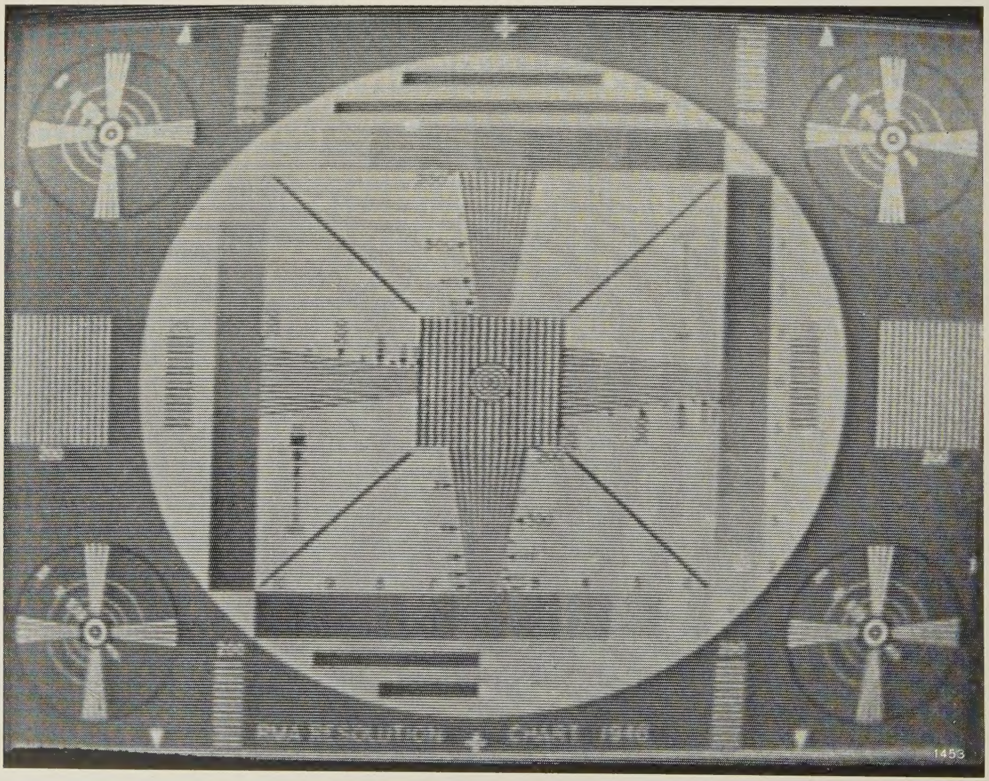

Fig. 6. Picture of a resolution chart picked up by a television camera, stored on a magnetic wheel and displayed on a monitor. For this recording, use was made of frequency modulation as in fig. 4. The raster consisted of over 300 lines.

wavelength was taken to be 8 μ. It therefore follows from $v_s = \lambda f$ that the peripheral speed of the wheel must be 64 m/sec. To achieve this at 3000 r.p.m. the wheel must have a diameter of 40 cm.

One of the wheels used in numerous experiments is shown in *fig. 7*. Two heads are disposed around the rim: an erasing head and a head serving alternately for writing and reading. Neither of the heads must touch the wheel, otherwise rapid wear of the heads and of the magnetic coating would result. Provided the erasing current is strong enough, the distance d between erasing-head and rim need not be extremely small. The distance between the write-read head and the rim is critical, however, and must not exceed about 1 μ. In the reading process the following relation exists between the attenuation a of the signal, the distance d and the wavelength λ [8]):

$$a = 55 \frac{d}{\lambda} \text{ dB}.$$

For $\lambda = 8$ μ and $d = 1$ μ, a is as much as 7 dB. The writing process, which is difficult to express in a formula, is even more critical in this respect.

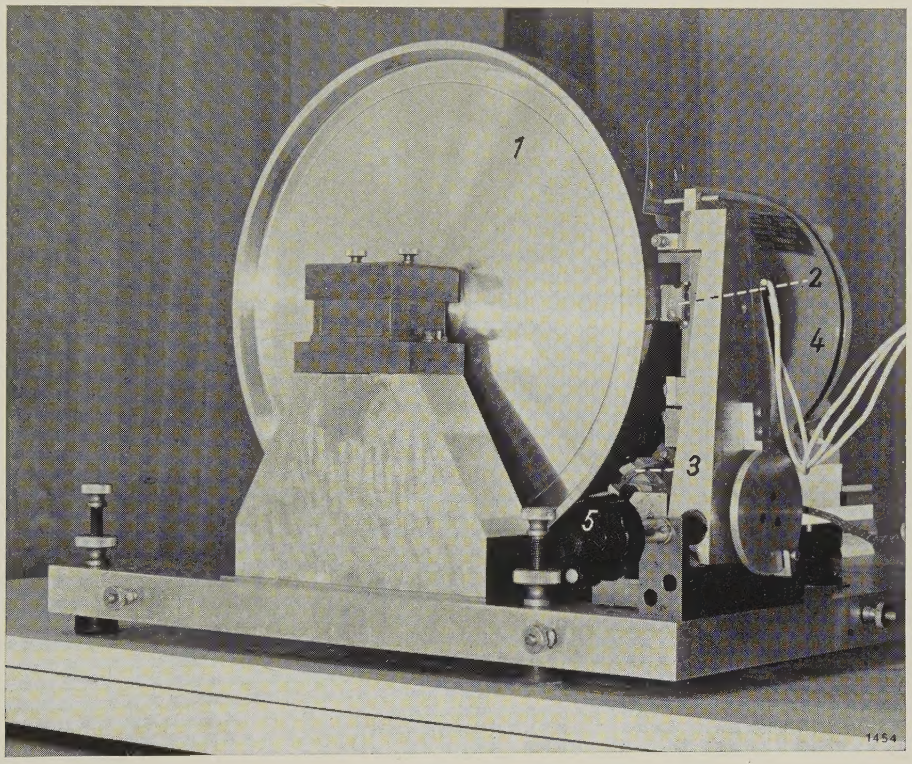

Fig. 7. *1* magnetic wheel. *2* write-read head. *3* erasing head. *4* motor. *5* knob for axially displacing the heads (the 30 mm wide rim flange can accommodate a large number of tracks side by side).

The mechanical construction therefore calls for the highest precision. As can be seen in fig. 7, the rim of the wheel has the form of a flange, 30 mm in width; without this the wheel would suffer too much deformation at high speeds. The magnetic layer is applied over the whole width of the flange, thus providing space for numerous tracks side by side. The wheel is mounted on a thick shaft (to withstand bending moments) and is machined in its bearings (journal bearings of exceptionally high quality) to within a tolerance of better than 1 μ [9]).

Since temperature variations may easily cause a change of a few microns in the radius, the write-read head cannot be rigidly mounted in a fixed position, otherwise, with a change in temperature, it would either touch the wheel or be too far away from it. In the construction chosen the head is arranged to ride on an air cushion when the wheel is turning at the right speed. The head *2* (*fig. 8*) is attached to an arm *6* which pivots about the point *7*. The air dragged round with the wheel pushes the head outwards against the pressure of a spring *8*, thus ensuring that a certain distance is maintained between the head and the turning wheel.

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[8]) See e.g. H. G. M. Spratt, Magnetic tape recording, Heywood, London 1958, p. 84.

[9]) This precision work was done very skilfully by L. M. Leblans of this laboratory.

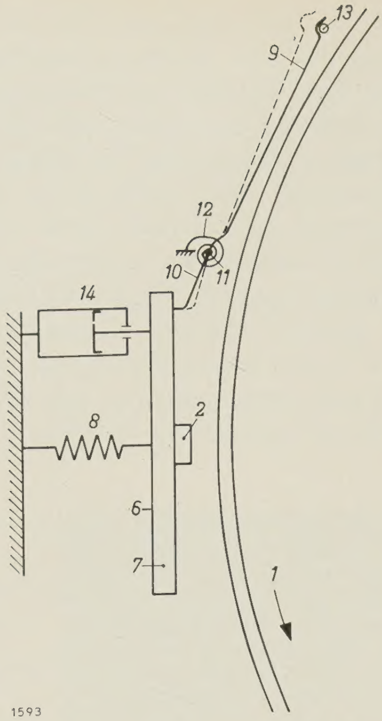

Fig. 8. Illustrating the method of maintaining a constant distance of about 1 μ between the magnetic wheel *1* and the writing-reading head *2*. *6* arm with pivot *7*. *8* compression spring. *9* vane, with arm *10* and pivot *11*. When the wheel is stationary the coil spring *12* forces the vane against the stop *13*; the arm *10* then prevents the spring *8* from pressing the head against the wheel. When the wheel is turning at full speed the air stream around the wheel pushes the vane *9* outwards (to position of dashed line) against the action of the coil spring *12*; arm *10* then releases arm *6*, and spring *8* pushes the head towards the wheel. The air film dragged round by the wheel forms an air cushion on which the head rides at a constant distance of about 1 μ from the wheel. *14* oil damping.

An optical test, made by passing light through the gap, demonstrated with a high degree of probability that the distance *d* remains roughly 1 μ. Oil damping *14* protects the head from vibrations, without preventing it from following slow changes due to temperature variations.

Special measures are needed to prevent the spring *8* from pressing the head against the wheel when the wheel is stationary, not turning fast enough or slowing down to a standstill [10]. For this purpose, use is made of a vane *9* connected to an arm *10*, pivoting about point *11* (see fig. 8). If the wheel stops or turns too slowly, a coil spring *12* holds the vane *9* against a stop *13*, and the arm *10* prevents arm *6* from moving the head into contact with the wheel. When the wheel comes up to full speed, the dragged air current pushes the vane *9* back and arm *10* releases arm *6* (dashed line in figure), thus restoring the above-mentioned equilibrium between the force exerted on the head *2* by spring *8* and that exerted by the air current.

Circuit for writing and reading a single frame

To write a single frame on the wheel, the writing head must be supplied with the frequency-modulated television signal for the duration of one frame ($^1/_{50}$ sec). This calls for a circuit that will keep the recording amplifier preceding the write-read head normally blocked but will unblock it during the first frame to begin after the depression of a push-button.

The block diagram of this arrangement is shown in *fig. 9*. A_2 is the recording-head amplifier. F_1 and F_2 are flip-flops, each with two stable states, *I* and *II*. Both receive at their inputs, *1*, a continuous train of (positive) frame-synchronizing pulses — sync pulses — which are separated in the usual way from the video signal in the circuit *S*. These sync pulses, represented in *fig. 10a*, keep the two flip-flops in state *I*, whilst F_2 delivers a biasing voltage which blocks the amplifier A_2. When the button *B* is depressed the discharge of capacitor *C* causes a pulse (fig. 10*b*) to appear at the input *2* of F_1; this pulse brings F_2 into state *II*. The first sync pulse now to arrive returns F_1 to state *I* (fig. 10*c*) which, via the coupling between F_1 and input *2* of F_2, has

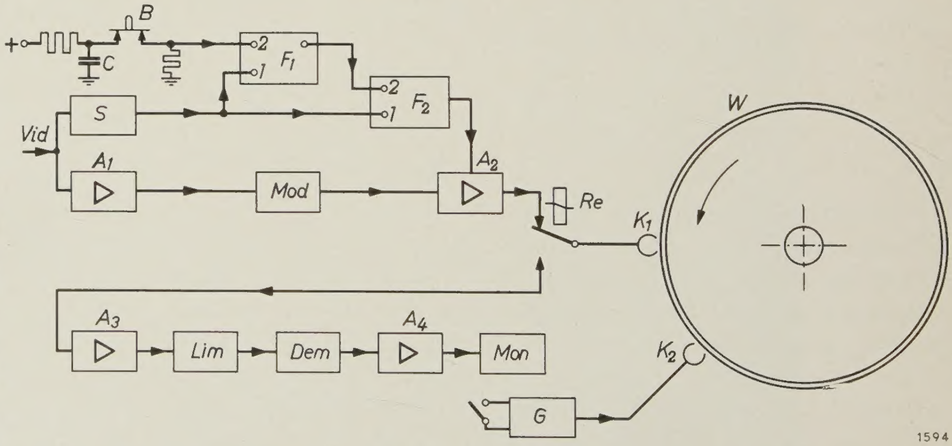

Fig. 9. Block diagram of the circuit for writing and reading a single frame. *Vid* video-signal input. A_1 video amplifier. *Mod* modulator, in which the video signal modulates a carrier wave in frequency. A_2 recording amplifier. *Re* write-read relay. K_1 write-read head. *W* wheel store. *S* circuit which separates picture-synchronizing signals — "sync pulses" — from video signal. F_1 and F_2 bistable flip-flops, with inputs *1* and *2*. *B* push-button, which, by discharging capacitor *C*, produces the "display" pulse. A_3 read amplifier. *Lim* limiter. *Dem* demodulator (frequency detector). A_4 video amplifier. *Mon* monitor. K_2 erasing head. *G* erasing-current generator.

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[10] The solution found for this problem was devised by J. F. van Oort of the Philips Exhibition Department.

the effect of causing F_2 to change from *I* to *II* (fig. 10*d*); the next sync pulse to arrive returns F_2 to state *I*. During exactly one frame, then, F_2 is in state *II*, and the recording amplifier is opened. This single frame, the first to follow the depression of the button, is therefore recorded on the wheel.

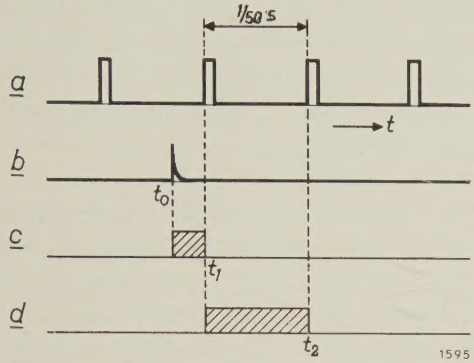

Fig. 10. *a*) Pulse train at the inputs *1* of flip-flops F_1 and F_2 in fig. 9. *b*) "Display" pulse at input *2* of F_1, produced when button *B* is depressed. *c*) During the interval t_0-t_1, F_1 is in state *II*. *d*) During the interval t_1-t_2 (= duration of one frame) F_2 is in state *II*. The recording amplifier A_2 (fig. 9) is then unblocked and the picture is recorded on the wheel.

Synchronizing the wheel with the video signal

An important application of the magnetic wheel store is that in which the recorded picture originates from a television camera at the same location. This is the case in the radiological application presently to be discussed. The frame frequency of the camera and the monitor, and the speed of revolution of the wheel, can then be governed by the frequency of the local electricity mains. As far as the wheel is concerned, this amounts to the use of a synchronous motor.

Cases also arise, however, where the frame frequency of the television picture to be recorded is not exactly equal to that of the local mains. Steps must then be taken to synchronize the wheel with the given frame frequency. The system which we have devised for this purpose consists of a flexible eddy-current coupling between the (non-synchronous) motor and the wheel, in combination with an electrical control system for governing the speed of revolution of the wheel.

Eddy-current coupling

Fig. 11 shows two cross-sections of the eddy-current coupling, which consists of the following components:

a) a flanged aluminium pulley *1*, at the rim of which is fitted a hollow steel cylinder *2* with copper lining *3*;

b) a sectored rotor *4*, keyed to the shaft *5* that drives the magnetic wheel;

c) a steel housing *6* containing a field coil *7*.

In relation to the pulley *1* and the housing *6*, the shaft *5* can rotate freely in ball-bearings. The pulley is motor-driven by means of a belt; the housing and the field coil are stationary. A variable direct current is passed through the field coil. This excitation current produces a magnetic flux through the components *6*, *4* and *2*, which concentrates in the sectors of *4*. When the cylinder *2* rotates and the rotor *4* is still stationary, eddy-currents are induced in the copper lining *3*. As a result, a torque is exerted on the rotor, causing the rotor — and hence the magnetic wheel — to rotate in the same sense as the pulley, but with a smaller angular velocity. When the angular-velocity difference $\omega_1 - \omega_2$ is small, the torque is proportional to $\omega_1 - \omega_2$ (*fig. 12*); for large differences in angular velocity, a phase shift exists between the pulsating magnetic field produced by the sectors and the induced eddy currents, causing the curve to bend over and a maximum to appear. When the wheel is started up, ω_2 is initially zero, hence $\omega_1 - \omega_2 = \omega_1$; if the maximum lies approximately at this value, the starting torque is high and the wheel quickly reaches full speed.

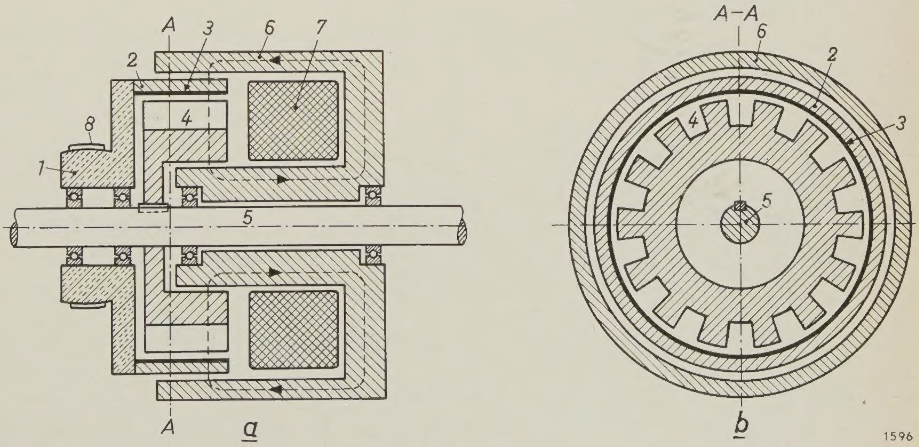

Fig. 11. Eddy-current coupling. *a*) axial cross-section, *b*) transverse section through *A-A* in (*a*). *1* flanged aluminium pulley with steel cylinder *2* and copper lining *3*. *4* sectored rotor, keyed to shaft *5* which, via a flexible coupling, drives the shaft (not shown) of the magnetic wheel. *6* steel housing. *7* field coil. *8* driving belt.

The control system

Synchronism between the speed of revolution of the wheel and the frame frequency of the video signal is obtained by causing any deviation from synchronism to react on the excitation current of the coupling. This is done by comparing the phase of two pulse trains: the sync pulses and wheel pulses.

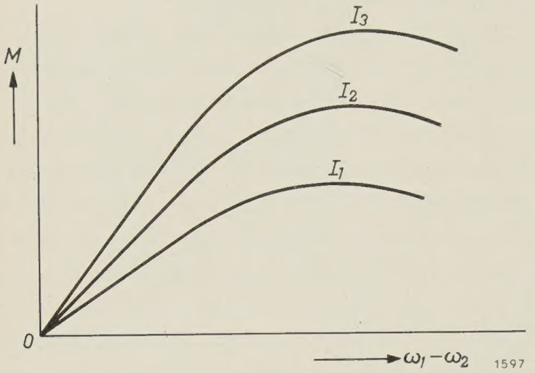

Fig. 12. Torque M, transmitted by eddy-current coupling, as a function of the angular-velocity difference ω_1—ω_2 between the pulley *1* and the shaft *5* (in fig. 11), for three values of exciting current ($I_3 > I_2 > I_1$).

As mentioned above, the sync pulses are derived from the video signal; the wheel pulses are induced in an appropriate pick-up head by a small magnet of ferroxdure fixed to the wheel. Any phase difference arising between the two pulse trains changes the excitation current, via a special circuit, in such a way as to make the phase difference smaller.

The circuit diagram of the control system is shown in *fig. 13.* The triodes T_1 and T_2 form part of a bistable flip-flop with a common cathode resistor R_2. The grid of the cathode follower T_3 is coupled to the anode of T_2. When the flip-flop is in state *1* (T_1 conducting, T_2 cut-off) the anode potential of T_2 is high and so too, therefore, is the grid potential of T_3 with respect to earth; T_3 therefore passes current, and thus the potential $v_{\rm k}$ across the cathode resistor R_3 is high ($v_{\rm k} = V_1$). In state *2,* on the other hand, $v_{\rm k}$ is low ($= V_2 < V_1$). When the potential v_{C1} across C_1 drops below a certain critical value $V_{1,2}$, the flip-flop changes from state *1* to state *2.* For the change from *2* to *1,* v_{C1} must exceed another critical value, $V_{2,1}$. The levels of $V_{1,2}$ and $V_{2,1}$ in relation to V_1 and V_2 are indicated in *fig. 14.*

Roughly, the circuit functions as follows. When synchronism is approximately achieved the flip-flop is brought into state *1* by every (positive) wheel pulse, and into state *2* by every (negative) sync pulse. The potential $v_{\rm k}$ of the cathode K is thus alternately high and low, remaining longer high the longer state *1* lasts, i.e. the more time it takes before a wheel pulse follows a sync pulse.

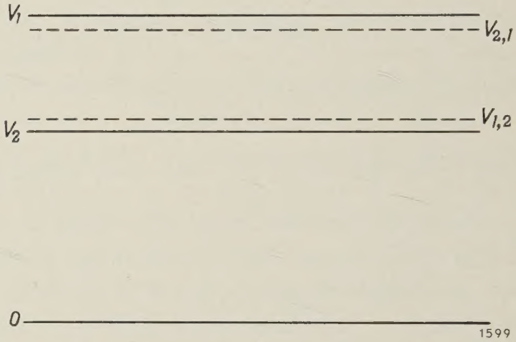

Fig. 14. V_1 potential of point K (fig. 13) in stable state *1;* V_2 *idem* in state *2.* $V_{1,2}$ value which v_{C1} must reach to cause state *1* to change to state *2;* $V_{2,1}$ *idem* for transition from *2* to *1.*

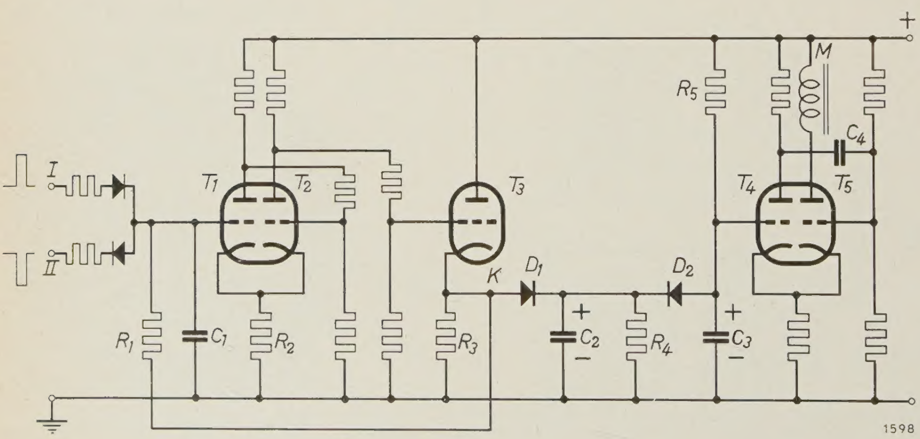

Fig. 13. Circuit diagram of control system. *I* input for (positive) wheel pulses. *II* input for (negative) sync pulses. *M* field coil of eddy-current coupling (*7* in fig. 11). If the wheel turns too slowly, the circuit energizes the electromagnet more strongly, and if the wheel turns too fast, less strongly. For explanation, see text.

The high voltage V_1 of K appears each time across the capacitor C_2 via the diode D_1. This capacitor discharges gradually through the resistor R_4. The longer the interval τ between a sync pulse and a wheel pulse, the lower is the final value to which the voltage v_{C2} across C_2 decreases. This final value is taken over via an "overflow diode" D_2 by capacitor C_3 and determines the current that flows through the triode T_4. Between each two decreases in v_{C2} the voltage v_{C3} across C_3 is able to rise again slightly as C_3 is charged up via R_5.

Let us assume that the wheel is turning a little too slowly. The interval τ will then show a tendency to increase. The potential v_{C3} therefore drops, the current through T_4 decreases, and, since T_4 and T_5 have a common cathode resistor, the decrease in the the current through T_4 causes an increase in the current through T_5. The latter current energizes the electromagnet in the eddy-current coupling. In the case under consideration, then, the excitation current rises and so, too, does the torque exerted on the rotor, causing the wheel to turn faster. Conversely, if the wheel starts to turn too fast, the excitation current is reduced, resulting in a drop in speed.

The capacitor C_4 between the anode of T_4 and the grid of T_5 influences the frequency response of the system in such a way as to prevent instability occurring.

A difficulty in phase control systems is often the starting-up process, when the phase relation between the two pulse trains is completely irregular. The usual practice is to switch-off the control system before starting-up, and to work with maximum torque until an electrical tachometer shows that the right speed has been reached; only then is the control system put into operation. This precaution is unnecessary with the arrangement in fig. 13; the control system can be switched on right from the beginning and will always ensure that synchronism is reached.

Each pulse from the wheel causes the voltage v_{C1} to jump by $V_1 - V_2$, and each sync pulse causes an equal downward drop. When the wheel is started up, the wheel pulses occur at first much more slowly than the sync pulses. The waveform of v_{C1} is then as shown in *fig. 15a*. The flip-flop is in state *2* (T_1 cut-off, T_2 conducting) and remains for a while in that state, because v_{C1} cannot reach the critical value $V_{2,1}$ as long as the wheel is turning much too slowly. In this state, v_k v_{C2} and v_{C3} have the low value V_2, which corresponds to strong excitation of the eddy-current coupling.

When the wheel approaches the correct speed of revolution the flip-flop will change its state now and then, but only when a wheel pulse is followed very quickly by a sync pulse, as in the intervals t_1-t_2 and t_3-t_4 in fig. 15*b*. During these intervals state *1* predominates, and v_k and v_{C2} have the high value V_1. When v_k jumps back to the low value V_2, capacitor C_2 begins to discharge through R_4 and therefore v_{C2} starts gradually to drop. In the case shown in fig. 15*b*, v_{C2} again reaches the value V_2, so that v_{C3} — which follows the lowest values of v_{C2} — still retains the value V_2 and thus the eddy-current coupling remains strongly energized.

The wheel now begins to turn rather too fast. Consequently the wheel pulses become somewhat more frequent than the sync pulses (fig. 15*c*) and a state is soon reached where v_{C1} and v_k jump continuously to and fro between the values V_1 and V_2. The voltage v_{C2} continually jumps up again before it can drop to V_2, and v_{C3} follows the successively somewhat

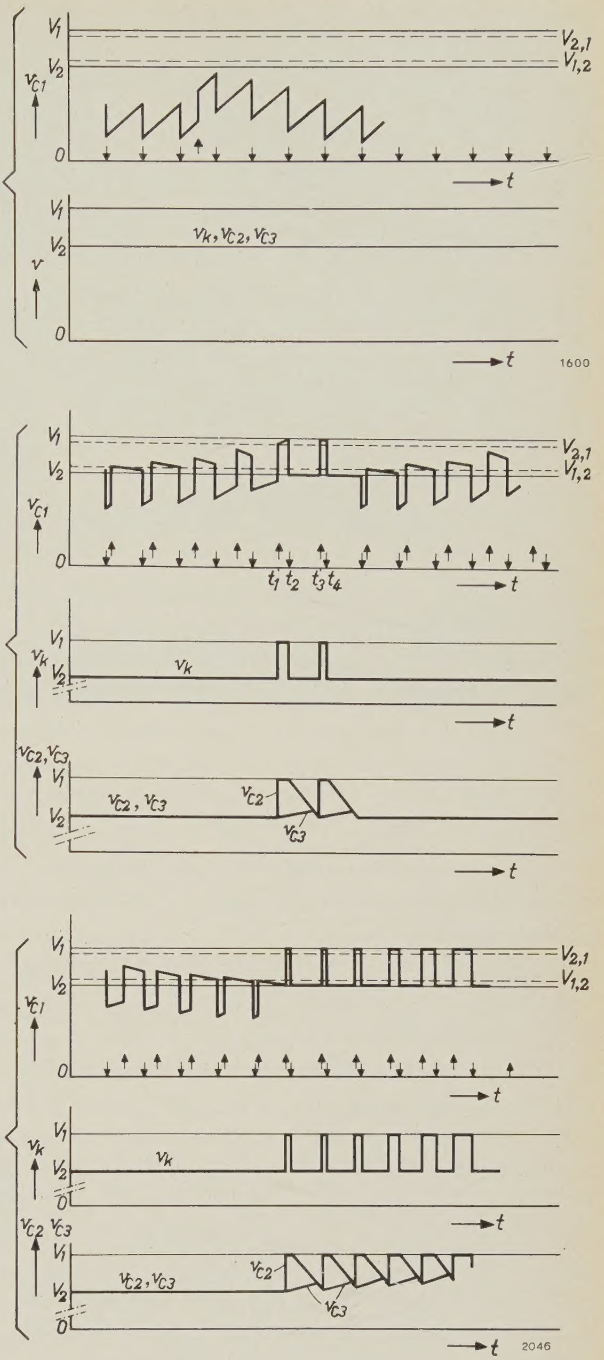

Fig. 15. Waveform of voltages v_{C1}, v_k, v_{C2} and v_{C3} in the control circuit (fig. 13). The arrows ↓ denote sync pulses, arrows ↑ wheel pulses.
a) The wheel has just been started and is turning much too slowly. The wheel pulses are consequently much less frequent than the sync pulses. v_{C1} does not attain the critical value $V_{2,1}$. The flip-flop T_1-T_2 therefore remains in state *2*; v_k, v_{C2} and v_{C3} are constant ($= V_2$), and the eddy-current coupling is strongly energized.
b) The wheel still turns rather too slowly. At $t = t_1$, v_{C1} reaches the critical value $V_{2,1}$, so that T_1-T_2 changes to state *1*, but the sync pulse at t_2 restores state *2*. The same happens at t_3-t_4. From t_1 to t_2 and from t_3 to t_4, $v_k = V_1$ and $v_{C1} = V_1$, but v_{C3} remains unchanged, and the coupling stays strongly energized.
c) The wheel begins to turn rather too fast, and locks back into synchronism. v_{C1} and v_k jump to and fro between V_2 and V_1; v_{C2} no longer drops to the level V_2; v_{C3} gradually rises and the coupling becomes less strongly energized, until the steady state is reached at synchronism.

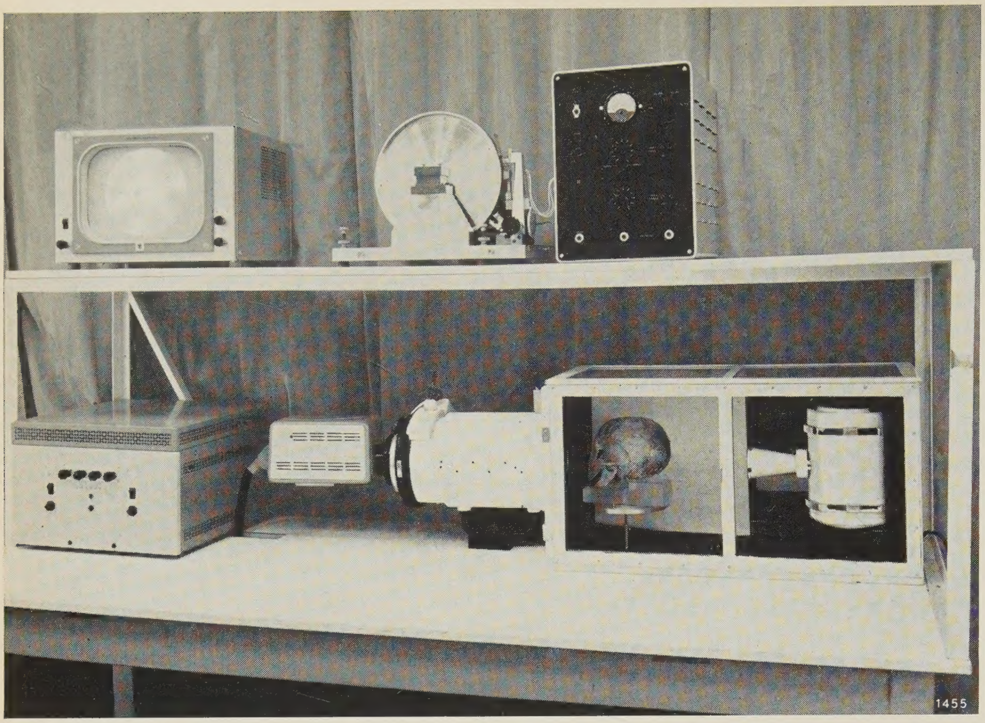

Fig. 16. Equipment for demonstrating the radiological application of the magnetic wheel store (as yet without frequency modulation). From right to left, below: X-ray tube, subject (a skull), X-ray image intensifier, television camera, auxiliary apparatus for television system; above: amplifying apparatus for recording and reproduction, magnetic wheel, monitor.

higher minima of v_{C_2}. (The charge needed to effect this increase in v_{C_3} is supplied to C_3 through R_5.) As v_{C_3} rises it increases the current through T_4 and decreases the current through T_5. The eddy-current coupling is therefore less strongly energized, the wheel turns more slowly and enters into synchronism.

Another virtue of the system is its relative insensitivity to changes in the frequency of the sync pulses; this is a consequence of the fact that a speed measurement (with an electrical tachometer) is not necessary.

Application in radiology

In the system described, as in all other magnetic recording systems, the recording can be preserved indefinitely or it can be erased; after erasure the magnetic layer can be immediately used again for a fresh recording. Since there is no contact between head and wheel, the recording can be reproduced as often as required. As we have seen, the picture quality is highly satisfactory when use is made of frequency modulation.

These features have led to a promising application fo the magnetic wheel store in radiology [11]). A picture of the image on an X-ray screen is taken with a television camera and one frame of the picture is recorded on the magnetic wheel. The X-ray image can then be displayed immediately on a monitor tube and viewed by the doctor as long and as often as necessary. In this way no time is lost in developing an X-ray film, and precious minutes can be saved during a surgical operation. The X-ray dose required is extremely low — very much lower than in a fluoroscopic examination of a few seconds, and, if an X-ray image intensifier is used, even lower than that needed for a radiograph. A further important advantage is that an unsatisfactory exposure can immediately be retaken.

Fig. 16 shows a photograph of the equipment used to demonstrate this application (as yet without frequency modulation) at the International Radiological Congress at Munich.

Summary. A magnetic wheel store is described on which a single television frame can be recorded. The picture quality obtained by direct recording of the video signal is quite serviceable for some purposes, in spite of the wide frequency range (50 c/s to 5 Mc/s). Considerable improvement results if the video signal is made to frequency-modulate a carrier and this latter recorded. The signal then recorded can have e.g. a frequency of 7 Mc/s in the white, 5.5 Mc/s in the black, and 5 Mc/s at the peaks of the sync signals, giving a frequency range from about 0.5 Mc/s to 8 Mc/s. In one experimental version the minimum wavelength impressed on the wheel is 8 μ at 8 Mc/s, the peripheral speed is 64 m/sec, the speed of revolution is 3000 r.p.m. and the wheel has a diameter of 40 cm. The rim of the wheel is 30 mm wide and can accommodate numerous tracks side by side. A device is described in which the air film dragged round with the wheel keeps the write-read head at a distance of about 1 μ from the magnetic coating of the wheel, without any contact. Also discussed are a circuit for writing and reading a single frame and a system of synchronizing the wheel with the video signal. Finally, mention is made of a promising application in radiology; advantages over photography are that it dispenses with the need for developing X-ray films, minimizes the X-ray dose and allows the immediate retake of unsatisfactory exposures.

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[11]) Th. G. Schut and W. J. Oosterkamp, The application of electronic memories in radiology, Medicamundi **5**, 85-88, 1959 (No. 3/4); Th. G. Schut and W. J. Oosterkamp, Die Anwendung elektronischer Gedächtnisse in der Radiologie, Elektron. Rdsch. **14**, 19-20, 1960 (No. 1).

RESONANCE ISOLATORS FOR MILLIMETRE WAVES

by H. G. BELJERS. 621.372.852.223:621.318.134

In microwave equipment frequent use is made nowadays of non-reciprocal devices. Principal among these is the directional isolator, a device that passes waves in the one direction without significantly attenuating them, and attenuates them very strongly in the other [1]).

The so-called resonance isolator makes use of the presence in a rectangular waveguide of a rotating magnetic field at certain places. If a suitable magnetic material is fixed at these places, gyromagnetic resonance occurs in the one direction of propagation and not in the other. Since gyromagnetic resonance — often referred to briefly as magnetic resonance — is attended by losses, the waves propagated in the first-mentioned direction are strongly attenuated. By using material free or almost free of other kinds of losses, it is possible in this way to make a directional isolator. The millimetre-band types discussed in this article are all resonance isolators.

The operation of other kinds of directional isolator depends on the Faraday effect or on so-called field-displacement. These can also be employed in the millimetre bands. Those based on the Faraday effect involve a round section of waveguide and it is necessary to fit transition sections if they are to be incorporated in a system using rectangular waveguides.

Gyromagnetic resonance occurs when a static magnetic field is applied perpendicular both to the direction of propagation and to the magnetic field of the microwaves. The required field H is roughly proportional to the frequency f of the microwaves. Expressing H in A/m and f in Mc/s, we can write:

$$H \approx f/0.035 \,. \quad . \quad . \quad . \quad . \quad . \quad . \quad (1)$$

The exact value of H is also governed by the demagnetization, in other words by the shape of the piece of material. This appears from Kittel's formula:

$$f = 0.035 \sqrt{H + M(N_x - N_y)} \times \sqrt{H + M(N_z - N_y)} \,, \quad . \quad . \quad . \quad (2)$$

where M is the saturation magnetization and N_x, N_y and N_z are the demagnetizing factors. (The direction of propagation is the z direction, and the magnetic field is parallel to the y direction.) As can be seen, formula (2) is equivalent to (1) if $N_x = N_y = N_z$, that is if the magnetic material is spherical in shape.

A further consequence of demagnetization is that, to achieve minimum damping in the forward direction, the microwave magnetic field should as a rule be elliptically and not, as might be thought at first sight, circularly polarized. We shall return to this point presently.

From formula (1) we may deduce that to make resonance isolators for wavelengths in the 8.6 and 4.3 mm bands (frequencies of 35 Gc/s and 70 Gc/s), now coming increasingly into use, we should need magnetic fields of about 10^6 A/m (12 500 oersteds) and 2×10^6 A/m (25 000 oersteds) respectively. The properties of the materials at present available for permanent magnets do not, however, allow of generating fields as high as 2×10^6 A/m, and although a field of 10^6 A/m is possible, it entails an unmanageably large and heavy magnet.

Nevertheless, a much weaker external field may be used, or it may be dispensed with altogether, if the resonance is produced in a hard magnetic material like a crystal-oriented anisotropic ferrite, that is a ferrite in which the preferred directions of magnetization of the crystallites are aligned parallel to each other. The anisotropic ferrites used in this case all possess hexagonal crystal structure, with the c-axis as the preferred direction of magnetization. The electron spins take up their preferred alignment parallel to this axis. It costs a great deal more energy to magnetize such material in a direction other than that of the hexagonal axis. The stiffness with which the spin orientation is bound to this preferred direction is expressed in terms of the magnetic field — the anisotropy field — that would have to be applied to an isotropic material in order to bind the spins with the same stiffness to the direction of that field. It follows from this definition that the field H, which, according to eq. (1), gives rise to magnetic resonance at a certain frequency, is simply equal to the sum of the anisotropy field H_a and the external field H_u.

Plainly, then, the external field can be dispensed with entirely if H_a has exactly the required value. Although this is attractive from the design point of view, it is not without its disadvantages. In the first place, if there is no external field the material is not always completely saturated, that is a greater number of Weiss domains have a "wrong" orientation, and as a result the damping of the microwaves in the forward direction is increased, which is obviously undesirable. In the second place, unless precautions are taken, the operation of such an isolator can be ruined by an interfering external magnetic field, which reduces the magnetization or may even cause it to disappear altogether if the interfering field is stronger than the coercivity of the material (approximately 2×10^4 A/m for the materials at present in use).

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[1]) H. G. Beljers, The application of ferroxcube in unidirectional waveguides and its bearing on the principle of reciprocity, Philips tech. Rev. **18**, 158-166, 1956/57.

In this article we shall describe isolators with and without an external field.

A most suitable material in resonance isolators for wavelengths in the 8.6 mm region (the Q band) is *topotactically* oriented material [2]) of composition $Ba(Zn_{0.35}Mn^{II}_{0.15}Ti_{0.5})(Fe_{0.95}Mn^{III}_{0.05})_{11}O_{19}$. This has an anisotropy field of more than 85×10^4 A/m, which means that the external field need not exceed about 18×10^4 A/m. For wavelengths in the region of 4.3 mm (the V band) a material having a much higher anisotropy field is needed. Ferrites of this kind have recently been developed in the Philips Irvington Laboratory [3]), and one of them, which has an anisotropy field of about 188×10^4 A/m, is eminently suited for use in a resonance isolator for 4 mm waves.

The anisotropy field is given the value required for a particular application by substituting other atoms for a certain fraction of the Fe atoms in the base material — a barium-ferrite for the 8 mm band and a strontium-ferrite for the 4 mm band. Because the magnitude of the anisotropy field is thus, within certain limits, controllable, it is proposed to designate this type of material by the collective name *controlled uniaxial anisotropy ferrites*, abbreviated to *c.u.a.f.* The resistivity of the two materials is very high (> 10^7 Ωcm); the dielectric losses are negligible.

It should be noted that the permissible power transmission of resonance isolators is limited by the temperature increase caused by the energy absorbed in the ferrite. If the temperature exceeds a certain value, the damping in the inverse direction begins to drop appreciably. In the isolators for 8 mm waves the energy dissipated should not exceed an average of 1 W.

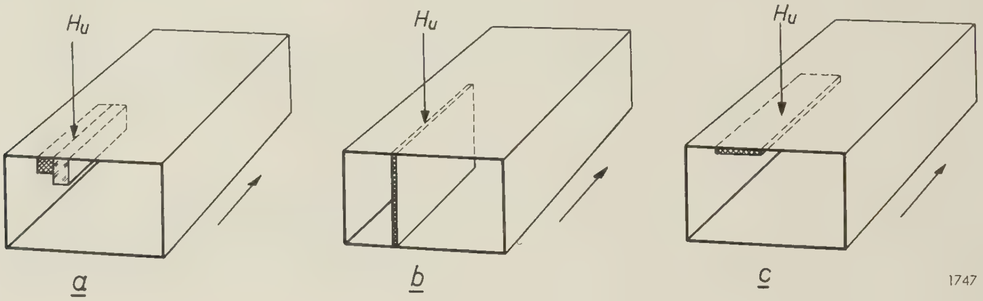

Fig. 1. Schematic representation of microwave resonance isolators in which the damping is produced by gyromagnetic resonance in a ferrite.
a) Actual arrangement
b) Extreme case, with thin ferrite sliver parallel to the plane of the electric lines of force (*E* plane).
c) Other extreme, with ferrite parallel to plane of magnetic lines of force (*H* plane).
In all three figures H_u is the external magnetic field, and the arrow at the right indicates the forward direction of propagation through the waveguide.

Isolators with weak external magnetic field

The construction of resonance isolators operating with an external field is illustrated schematically in *fig. 1a.* Fitted at a suitable place, side by side, on the broad wall of a rectangular waveguide are a square bar of c.u.a.f. material and a fused quartz strip. This form and the location of the ferrite, lie between two extreme cases where a very thin ferrite sliver is located either in a plane parallel to the electric lines of force (see *fig. 2*) or in a plane parallel to the magnetic lines of force (fig. 1*b* and *c*). In the first of these extreme cases the field in the waveguide is considerably distorted owing to the presence of the ferrite sliver, and measures are needed to minimize the resultant reflection. This is often done by making the ferrite sliver trapezium-shaped. Furthermore, the ratio of the attenuations (measured in decibels) undergone by the waves in the forward and inverse directions — this ratio may be regarded as a kind of figure of merit — is usually not so favourable as in an isolator in which the ferrite sliver

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[2]) See F. K. Lotgering, Topotactically crystal-oriented ferromagnetics, Philips tech. Rev. **20**, 354-356, 1958/59.

[3]) F. K. du Pré, D. J. de Bitetto and F. G. Brockman, Magnetic materials for use at high microwave frequencies (50-90 Gc/s), J. appl. Phys. **29**, 1127-1128, 1958.

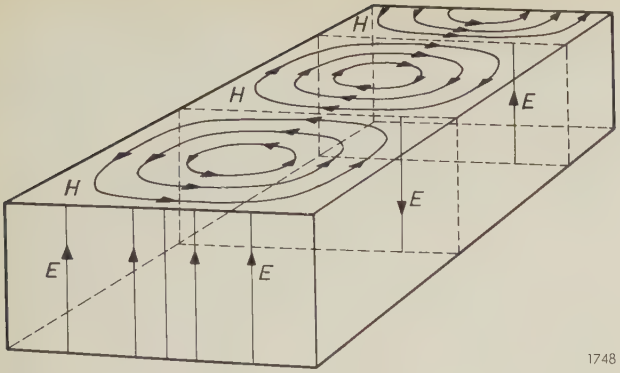

Fig. 2. In the TE_{10} mode of vibration of a rectangular waveguide the electric lines of force (E) are perpendicular to the broad side faces and the magnetic lines of force (H) are parallel to these faces.

is parallel to the magnetic lines of force. Isolators of the latter type, however (fig. 1*c*), demand a somewhat stronger magnetic field — which is disadvantageous only when a soft ferrite is used — and the width of the sliver, that determines the maximum attenuation that can be achieved per unit length [4]), is rather limited.

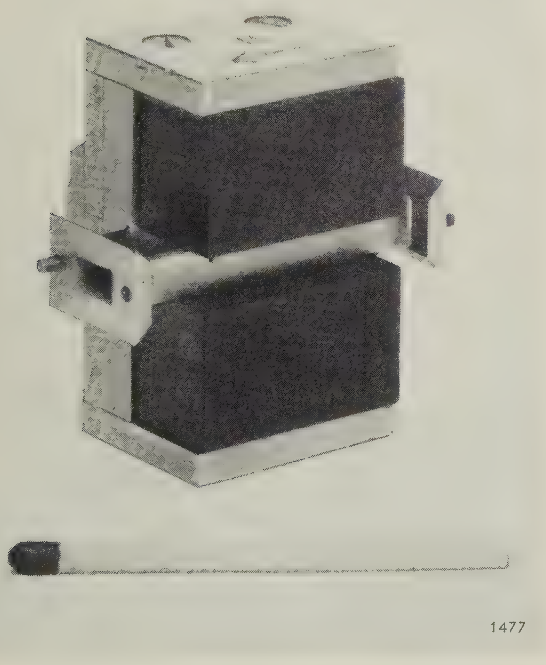

Fig. 3. Isolator for 8.6 mm waveband, using controlled uniaxial anisotropy ferrite material (anisotropy field 85 × 10^4 A/m) and a weak external magnetic field (18 × 10^4 A/m). This field is generated by a permanent magnet consisting of two blocks of ferroxdure in an iron yoke.

The intermediate form which we have chosen, and which has proved entirely satisfactory in practice, does not require oblique shaping of the ferrite. Because of the fairly considerable thickness of the ferrite bar, the attenuation per cm length is high. So, too, is the ratio of the attenuations in the forward and inverse directions. This favourable property is partly due to the presence of the silica strip. Since much of the microwave energy traverses the waveguide via and close to the silica strip (dielectric constant ≈ 4, dielectric losses minimal), the attenuation in the inverse direction is substantially greater than in an isolator without dielectric, but the attenuation in the forward direction is not. The explanation of these effects is complicated and not yet wholly clear.

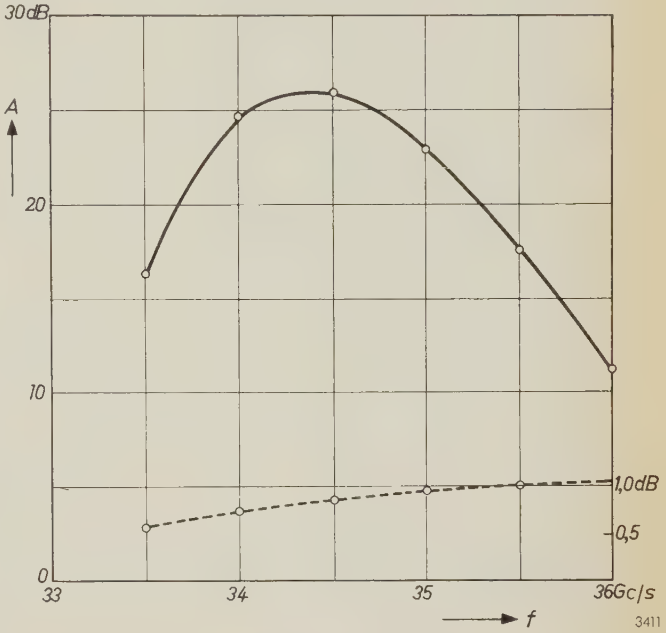

Fig. 4. Transmission characteristics (attenuation A versus frequency f) of the resonance isolator with external magnetic field, for the 8.6 mm wave band. The scale values on the left relate to the curve for the inverse direction (solid line), those on the right to the curve for the forward direction (broken line). The maximum ratio between the attenuations (in decibels) occurring in the two directions is approximately 30, and is obtained at a frequency of about 34.3 Gc/s ($\lambda = 8.7$ mm).

In the isolator for the 8-9 mm wave band [5]) (*fig. 3*) two bars of c.u.a.f. material, having an anisotropy field of 85×10^4 A/m, are mounted end to end. Their dimensions are $12 \times 0.80 \times 0.40$ mm and $12 \times 0.60 \times 0.36$ mm. The slight difference in their widths makes it possible to obtain a broader characteristic (*fig. 4*), inasmuch as the resonance frequencies differ somewhat for the two bars owing to the slight disparity between their demagnetizing factors. The external magnetic field is roughly 18×10^4 A/m. It is provided by a permanent magnet consisting of two blocks of ferroxdure held in an iron yoke. An air gap, which can be bridged by a shunt allows fine adjustment of the external field strength; the latter is so adjusted that, together with the anisotropy field, it yields exactly the value required to produce gyromagnetic resonance at 35 Gc/s.

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[4]) The fact that a stronger magnetic field is needed may be deduced directly from formula (2). In the case of the H plane strip we have $N_y \approx 1$ and $N_x \approx N_z \approx 0$, whereas for the E plane: $N_x \approx 1$ and $N_y \approx N_z \approx 0$.

[5]) See also H. G. Beljers, Ferrite isolators in the 8-9 mm waveband, Commun. Congrès int. Circuits et Antennes Hyperfréquences, Paris 21-26 Oct. 1957, Part II (Suppl. Onde électrique **38**, No. 376 ter), pp. 647-648, 1958.

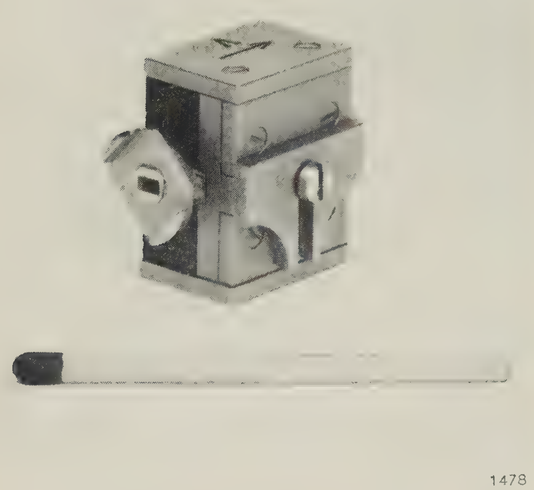


Fig. 5. Resonance isolator for 4.3 mm waves. Anisotropy field of magnetic material approx. 188×10^4 A/m. External magnetic field approx. 16×10^4 A/m. The movable shunt on the front serves for adjusting the external field to the exact value required.

The resonance isolator for the 4-5 mm waveband (*fig. 5*) uses bars of c.u.a.f. material which are half as large as in the 8.6 mm isolator and have an anisotropy field of 188×10^4 A/m.

Resonance isolator without an external magnetic field

The construction of the isolator about to be described, which operates without an external field, differs considerably from that of the other. This isolator uses two c.u.a.f. ferrite strips one on each side of a thin sliver of dielectric material (in this case aluminium oxide with a dielectric constant ε of 9). The two ferrite strips, which, like the dielectric, take up the whole height of the waveguide, are magnetized in opposite directions. Together with the walls of the waveguide, which are of iron to give magnetic screening, they form a closed magnetic circuit, that is to say there is no demagnetization. The plate with strips is mounted centrally in the waveguide (*figs. 6* and *7*).

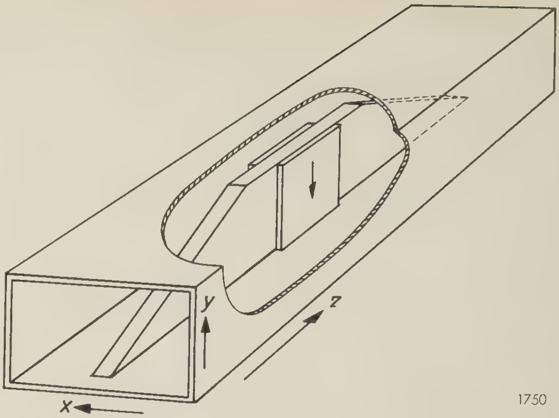


Fig. 6. Schematic representation of a resonance isolator for 8.6 mm waves, which needs no external field. Two thin ferrite strips ($2.0 \times 3.5 \times 0.15$ mm; anisotropy field 85×10^4 A/m) are fixed to the sides of a plate of aluminium oxide (thickness 1.1 mm) mounted centrally in the waveguide. The ends of the plate are cut obliquely to avoid reflections. The visible ferrite strip is magnetized in the direction of the arrow, the other in the opposite direction. The forward direction is that of the positive z axis.

Here, too, a large part of the microwave energy passes through and near the dielectric. Calculations show that the effect of the side walls of the waveguide is of secondary significance, and further that the magnetic field of the microwaves in the side faces of the dielectric plate is elliptically polarized. The ellipticity (by which is meant the ratio H_x/H_z,

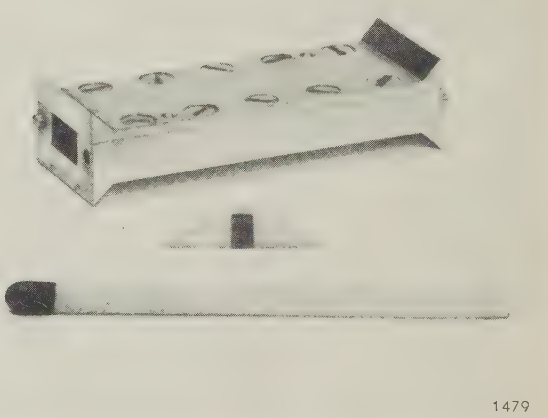


Fig. 7. Resonance isolator for 8.6 mm waves, with no external field, as schematically illustrated in fig. 6. In front of it can be seen the aluminium-oxide plate with ferrite strips, used in isolators of this type.

cf. fig. 6) decreases asymptotically with increasing plate thickness to the value $\sqrt{\varepsilon/(\varepsilon-1)}$, that is in our case to about 1.06. As mentioned earlier, to obtain minimum damping in the forward direction, elliptical polarization is precisely what is wanted. Calculation shows that the ellipticity of the rotating field must have the value

$$\sqrt{\frac{H_a + M(N_x - N_y)}{H_a + M(N_z - N_y)}}\,.$$

For very thin ferrite strips this expression approximates to

$$\sqrt{(H_a + M)/H_a}\,,$$

which in our case comes to about 1.12.

From the above we may infer that the thickness of the aluminium-oxide plate is here of great importance, and also that it is possible to choose this thickness such that the imposed requirements are fulfilled. Owing to the symmetrical arrangement of the whole assembly it is obviously not possible, as it was in the other two isolators discussed, to choose the dimensions of the materials more or less freely and then to minimize the damping in the forward direction by determining the most favourable position in the waveguide for the plate with the ferrite strips.

The best result is obtained when the dielectric and ferrite are given the dimensions indicated in the caption to fig. 6, which are in good agreement with the calculated values. To avoid reflections the plate is again cut obliquely at the ends. The same ferrite material can be used as in the isolator operating with an external field, the correct resonance frequency being obtained because of the entirely different shape of the ferrite strips. *Fig. 8* shows the characteristic of one of the first isolators of this type made during the development stage.

The follwing calculation will serve to demonstrate the fact that, in this isolator with no external field, the required resonance frequency is nevertheless obtained with the same c.u.a.f. material. For very thin ferrite samples N_y and N_z are negligible and $N_x \approx 1$, so that formula (2) can be written:

$$f = 0.035\sqrt{(H + M)H}\,. \quad . \; . \; . \; . \; . \quad (3)$$

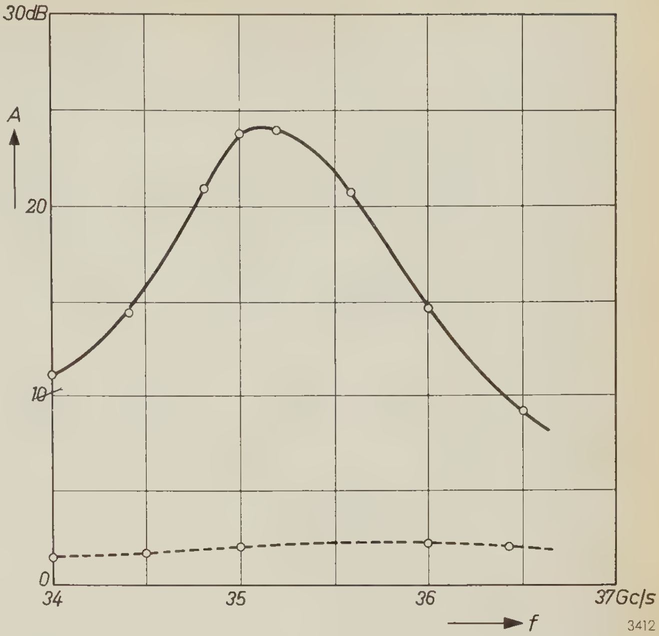


Fig. 8. Transmission characteristics of the resonance isolator without external field. The attenuation ratio here reaches a maximum of 12.

For a material with $M = 28\times10^4$ A/m (3500 gauss) and for a resonance frequency of 35 Gc/s, we obtain an H value of 86×10^4 A/m. This is more or less the strength of the anisotropy field. For the isolator with external field, where $N_x \approx {}^4/_5$, $N_y \approx {}^1/_3$ and $N_z \approx 0$, we have:

$$f = 0.035\sqrt{(H_a + H_u + {}^7/_{15}M)(H_a + H_u - {}^1/_3M)}\,, \quad (4)$$

which, with $H_u = 18\times10^4$ A/m, yields an H_a value of more than 82×10^4 A/m. (The fact that N_x is here ${}^4/_5$ and not ${}^2/_3$, as might be expected, is due to the fact that the image of the ferrite in the wall must be taken into account.) In view of the effect of the dielectric on the resonance frequency, the agreement may be described as reasonably good.

Summary. The gyromagnetic resonance effect, which occurs in magnetic materials in the presence of a suitable magnetic field, can be utilized for making nonreciprocal microwave transmission devices, such as directional isolators. In the millimetre wave bands the magnetic field strength required is extremely high (10^6 A/m for 8 mm waves and 2×10^6 A/m for 4 mm waves) and could not normally be generated with a permanent magnet of manageable proportions. Crystal-oriented anisotropic ferrites have now been developed, however, which possess in one direction a very high anisotropy field, and since the strength of this field can be deducted from the total magnetic field required, it is possible to use these materials for constructing isolators that need only a weak external field or none at all. The article describes resonance isolators of this type for wavelengths in the region of 8.6 mm and 4.3 mm (35 and 70 Gc/s, respectively).

APPLICATIONS OF MICROWAVE TRIODES

by J. P. M. GIELES.

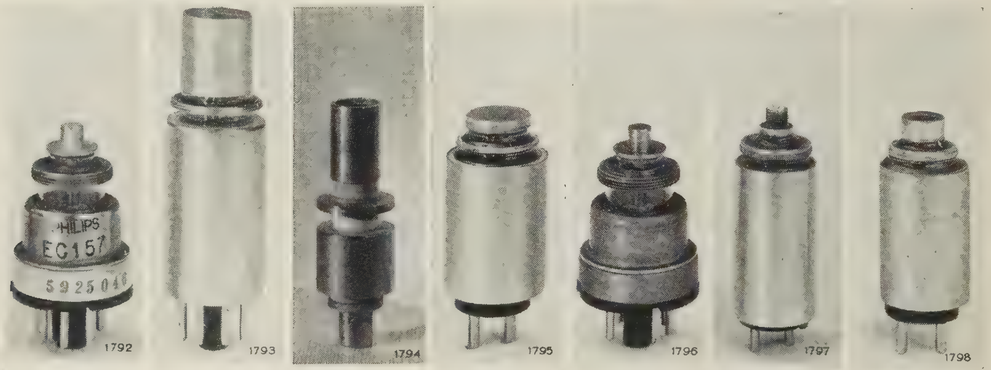
621.385.3.029.6

In recent years various disc-seal triodes have been developed at Philips which are capable of operating at wavelengths of 7.5 cm, 5 cm and less. The article below discusses a number of applications of these triodes that have already been realized, and others that are still in the development stage. Most of the applications are in microwave radio links for telephony and television.

Microwave triodes possess several attractive features. In addition to favourable phase characteristics, long life and easy replacement, they are easy to construct compared with other microwave tubes, they require no very high voltages and they can give a high $G \times B$ product, i.e. a considerable gain G even when the bandwidth B is large.

Another advantage of microwave triodes over other tubes used in the centimetre wavebands is their flexibility: triodes are relatively simple circuit elements and their simplicity makes them suitable for many and various functions. In complicated equipment containing many tubes, such as relay stations in microwave radio links, it is important for technical and economic reasons to limit the number of tube types to a minimum. It is therefore

Table I. Summary of data on microwave triodes manufactured or in development at Philips. The last two tubes are designed for a frequency of 6000 Mc/s, the others for 4000 Mc/s. For various purposes, however, the tubes can be operated up to frequencies about a factor of 2 higher.



	EC 157	EC 59	OZ 92	49 AL	22 EC	5 cm, 1 W	5 cm, 10 W
Literature references	1)2)7)8)9) 10)11)12)	3)4)7)9)12)	5)	12)	—	6)12)	12)
Cathode diameter (mm)	3.2	4.5	4.5	4.5	4	3.1	4.5
Cathode-grid spacing (μ)	40	60	60	40	40	25	40
Grid-anode spacing (μ)	240	300	300	300	240	300	500
Grid-wire diameter (μ)	7.5	30	30	15	12	7.5	15
Grid pitch (μ)	50	130	130	90	75	40	90
Anode voltage (V)	180	500	500	220	180	300	600
Anode current (mA)	60	250	250	200	140	60	300
Current density (A/cm²)	0.8	1.6	1.6	1.3	1.2	0.8	1.9
Transconductance (mA/V)	19	20	20	30	25	27	20
Low-level gain at 100 Mc/s bandwidth (3 dB below peak) (dB)	13	10	10	12	12	10	10
Output power at 8 dB gain (W)	1.5	12	12	6	5	1.2	—
AM-PM conversion at 8 dB gain (°/dB)	—0.8	—1.2	—1.2	—1.0	—	—	—
Cooling	air *)	water	water or air	air	air *)	air	water

*) An amplifier is in course of development in which cooling is effected by natural convection and radiation.

not surprising that the triode really comes into its own in such equipment, and has already found wide application in various installations built by Philips.

Table I gives a survey of the microwave triodes developed in recent years at Philips, or still in course of development. In this article we shall discuss in turn their application as:

amplifiers,
frequency multipliers,
oscillators,
mixers,
limiters,
amplitude modulators,
frequency modulators, and
high-level detectors.

Most applications have been studied on the EC 157. It is to be expected that the other types of microwave triode will behave similarly in equivalent circuits.

Amplification

Microwave triodes used for amplification purposes can be classified, according to signal level, as pre-amplifiers or as power amplifiers. Articles have appeared in this journal on the EC 157 and the 5 cm, 1 W tube as pre-amplifiers, and on the EC 59 and OZ 92 triodes as power amplifiers. For particulars, see the articles referred to in Table I. It should be noted that, in view of the high $G \times B$ product, the gain can be raised far beyond the tabulated values by reducing the bandwidth. For example, the bandwidth of an amplifier fitted with an EC 157 was reduced from 100 Mc/s to about 40 Mc/s, giving an increase in gain from 12 dB (= 16×) to 17 dB (= 50×). At a bandwidth of 40 Mc/s a cascade arrangement of three amplifiers with EC 157 triodes gave a total gain of 50 dB.

Not so well known is the application of the EC 157 as a *low-noise input stage* at frequencies below about 1000 Mc/s, used primarily in equipment for radio astronomy. Although its disc-seal construction (resulting in higher capacitances and limited switching possibilities) makes this tube less suitable as a pre-amplifier at frequencies far below the design frequency (4000 Mc/s), it is precisely there that the noise properties of the electrode system are favourable compared with those of a crystal mixer used as an input stage. The minimum noise figure, which is a tube characteristic, can be determined at any frequency by slightly varying the input matching. *Fig. 1* shows the average of the minimum noise figures, expressed in dB, of 13 type EC 157 triodes, measured as a function of frequency [13]).

The strongest point of microwave triodes, which is to produce simply and efficiently a high output power at a large bandwidth, makes them especially suited for use as *power amplifiers*. When used as such in microwave radio links it is most important that amplitude variations should not give rise to disturbing phase variations, and hence to distortion. The extent to which this occurs is called the

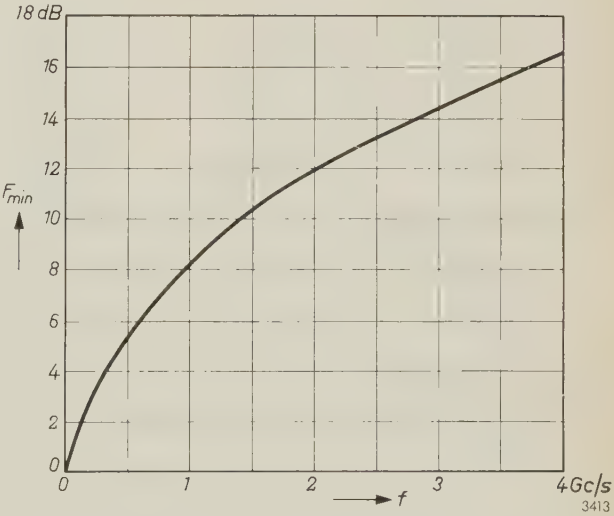


Fig. 1. Average $F_{\min}$ of the minimum noise figures in dB of 13 type EC 157 triodes, versus frequency.

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[1]) G. Diemer, K. Rodenhuis and J. G. van Wijngaarden, The EC 57, a disc-seal microwave triode with L cathode, Philips tech. Rev. **18**, 317-324, 1956/57. The EC 157 differs from the older type, EC 57, in having a cathode of longer life.

[2]) J. P. M. Gieles, A 4000 Mc/s wide-band amplifier using a disc-seal triode, Philips tech. Rev. **19**, 145-156, 1957/58.

[3]) V. V. Schwab and J. G. van Wijngaarden, The EC 59, a transmitting triode with 10 W output at 4000 Mc/s, Philips tech. Rev. **20**, 225-233, 1958/59.

[4]) J. P. M. Gieles and G. Andrieux, A wide-band triode amplifier with an output of 10 W at 4000 Mc/s, Philips tech. Rev. **21**, 41-46, 1959/60 (No. 2).

[5]) E. Mentzel and H. Stietzel, A metallic-ceramic disc-seal triode for frequencies up to 6000 Mc/s, Philips tech. Rev. **21**, 104-109, 1959/60 (No. 3).

[6]) M. T. Vlaardingerbroek, An experimental disc-seal triode for 6000 Mc/s, Philips tech. Rev. **21**, 167-171, 1959/60 (No. 6).

[7]) J. G. van Wijngaarden, Possibilities with disc-seal triodes, Onde électr. **36**, 888-892, 1956.

[8]) K. Rodenhuis, A 4000 Mc/s triode with L-cathode construction and circuit, Le Vide **12**, 23-31, 1957.

[9]) G. Andrieux, Amplificateurs de puissance à triodes pour 4000 Mc/s, Onde électr. **37**, 777-780, 1957.

[10]) J. P. M. Gieles, The measurement of group delay in triode amplifiers at 4000 Mc/s, Onde électr. **37**, 781-788, 1957.

[11]) M. T. Vlaardingerbroek, Measurement of the active admittances of a triode at 4 Gc/s, Proc. Instn. Electr. Engrs. **105 B**, Suppl. No. 10, 563-566, 1958.

[12]) H. Groendijk, Microwave triodes, Proc. Instn. Electr. Engrs. **105 B**, Suppl. No. 10, 577-582, 1958.

[13]) These measurements were done by G. A. W. J. Spanhoff and N. van Hurck. For the application of the EC 157 in an amplifier for radio astronomy, see C. L. Seeger, F. L. H. M. Stumpers and N. van Hurck, A 75 cm receiver for radio astronomy and some observational results, Philips tech. Rev. **21**, 317-333, 1959/60 (No. 11).

amplitude-modulation/phase-modulation (AM-PM) conversion of the tube; it is expressed in degrees of phase variation of the output voltage due to a variation of 1 dB in the input signal. The AM-PM conversion of triodes is of the order of $-1°$/dB, which is much less than that of most other short-wave tubes.

Special attention may be paid to the *parallel amplifier*. It is possible to connect two amplifiers in parallel such that, for the same gain, they deliver twice the output power. *Fig. 2* shows a parallel output stage of this kind for operation at 4000 Mc/s, equipped with two type 49 AL triodes. This stage is capable of delivering an output of 10 W at an anode voltage of 220 V, the gain being more than 8 dB. Both amplifiers, like a single amplifier, are equipped with ferrite isolators. (Because of this arrangement the data for fig. 2 differ somewhat from those given in Table I.)

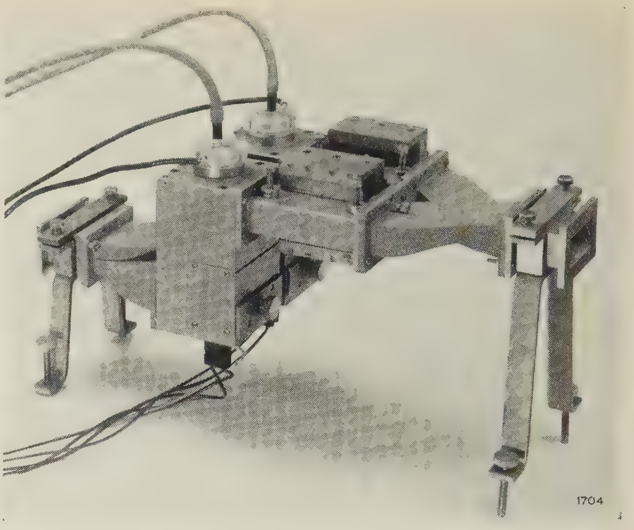

Fig. 2. Parallel output stage, consisting of two amplifiers each equipped with a type 49 AL triode.

Frequency	4000 Mc/s	Bandwidth	100 Mc/s
Anode voltage	220 V	Gain	8.2 dB
Anode current	2×200 mA	Output power	10 W
Cooling air:	27 l/min under pressure of 10 cm water.		

In this way outputs up to 50 W at 4000 Mc/s have been obtained from EC 59 triodes in the laboratory. Microwave radio links generally require no more than 10 or 20 W, which is ample to establish a dependable link. Apart from doubling the output power, the parallel output stage is more reliable than a single stage since, with proper design, the failure of one tube does not necessarily involve interruption of the link.

The use of triodes as microwave amplifiers may best be illustrated, perhaps, by their application in a "*straight-through*" *relay station* in a microwave radio link, i.e. a relay station in which the gain is obtained without demodulating to an intermediate frequency. This method of amplification stems from the fact that the triode as an amplifier of centimetre waves is superior to the tubes normally used for amplification in the IF wavebands. In view of switching and modulation problems a straight-through relay station does not lend itself directly to general use, but for long-distance links it has the great advantage of causing much less phase-distortion than conventional IF amplifiers. A laboratory version of such a straight-through relay station is shown in *fig. 3*. The block diagram is given in *fig. 4*.

A six-cavity input filter F_1 is followed by four pre-amplifier stages fitted with EC 157 triodes. All amplifiers are coupled by isolators D. By means of a frequency-shifting stage M, which will be discussed below under the head "Mixing", the frequencies of the input and output signal are separated sufficiently to exclude the possibility of spurious oscillations due to aerial feedback. On the recommendation of the Comité Consultatif International des Radiocommunications, the frequency shift is set at 213 Mc/s. After a three-cavity filter F_2 for suppressing the unwanted frequencies of the frequency-shifting stage, and three further amplifying stages, a small portion of the signal is taken off for automatic gain control. This feeds back through a DC amplifier AVC on the field current in the ferrite attenuator At, whose attenuation depends on the magnetic field produced by this current. This system ensures that the signal level at the input of the last stage remains constant within 0.5 dB for a variation of up to 25 dB in the input signal of the first stage. A single output stage is used, fitted with an EC 157, so that the output power of the whole installation is about 1.5 W. The overall bandwidth is about 40 Mc/s at 0.1 dB below peak, and the group-delay variation over 20 Mc/s amounts, without compensation, to about 1 millimicrosecond, which is less than 1/10th of the value in normal IF amplifiers. The intermodulation noise, which is due to the non-linearity of phase characteristics, is accordingly very low. The weak point of this relay station is the input noise factor, which, owing to the attenuation caused by the input isolator and the input filter, amounts to as much as 18 dB. This drawback can be overcome, however, by increasing the input signal, that is to say by raising the transmitted power of each relay station in the chain. If, for example, the parallel output stage shown in fig. 2, or a single output stage with an EC 59, be connected at the end of the arrangement described, an output power of 10 W can be achieved.

Finally, microwave triodes can also be used successfully as output amplifiers *at lower frequencies*. Wide use is made of the EC 157 in microwave links at frequencies in the 2000 Mc/s band. An experimental push-pull output stage with two EC 59 triodes in class C delivered an output of 76 W at 900 Mc/s [14]), which roughly corresponds to the performance of other tubes in that frequency band.

Frequency multiplication

The usual practice in microwave radio links is for the incoming radio-frequency signal to be transposed to the intermediate frequency by mixing it in a crystal mixer stage with the signal from a local oscillator. After amplification the IF signal is returned to the radio-frequency band by means of a second local oscillator (see section on mixing). The two RF local oscillators needed must be

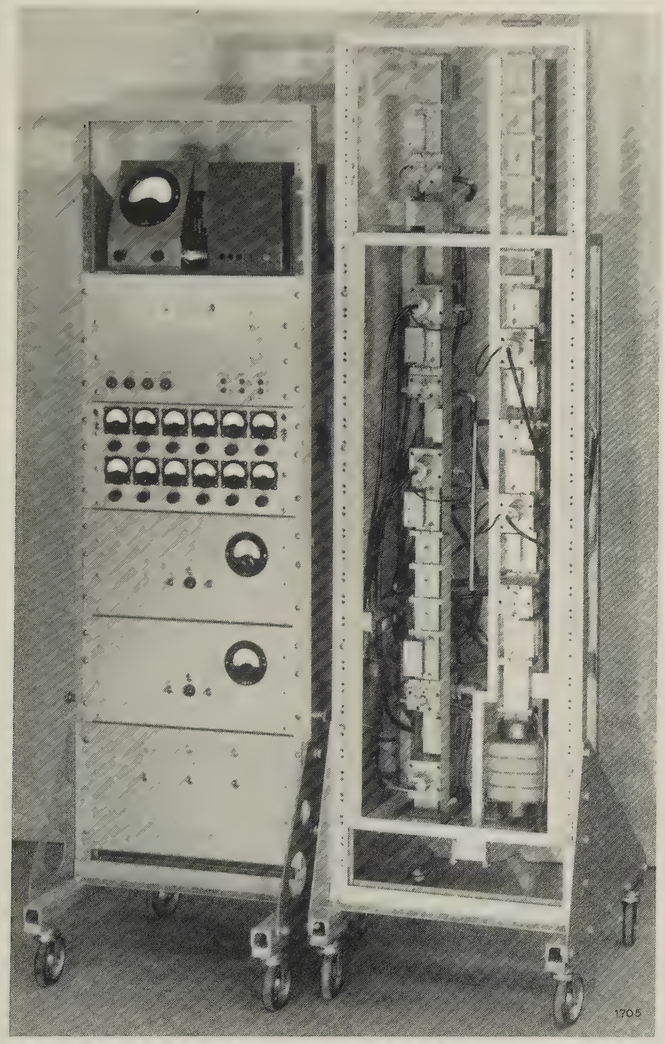

Fig. 3

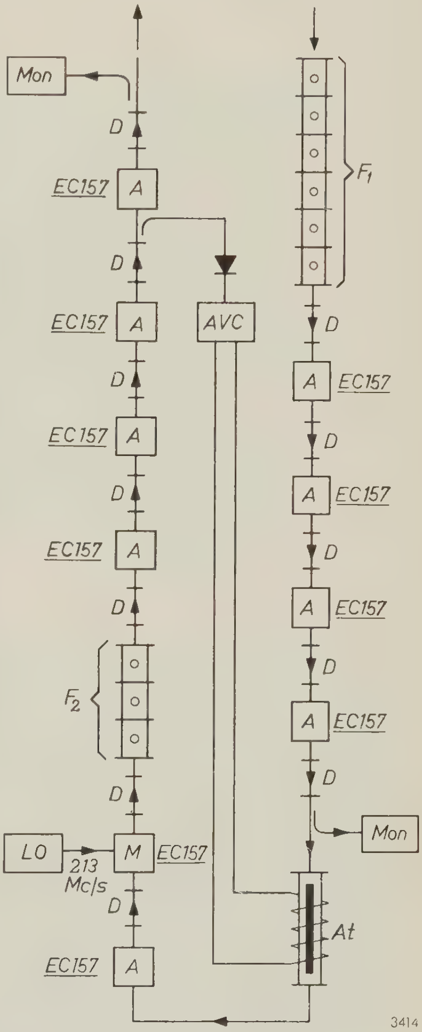

Fig. 4

Fig. 3. A complete "straight-through" relay station for a microwave radio link operating on 7.5 cm.

Fig. 4. Block diagram of "straight-through" relay station, in which all amplification is obtained in the 4000 Mc/s frequency band. F_1 six-cavity input filter. *D* ferrite isolators. *A* amplifier stages equipped with EC 157 triodes. *Mon* monitors. *At* ferrite attenuator. *M* mixer stage producing frequency shift between input and output signals. This frequency shift (213 Mc/s) prevents spurious oscillation due to aerial feedback. *LO* local oscillator. F_2 three-cavity filter for suppressing unwanted frequencies. *AVC* DC amplifier supplying the signal for the automatic gain control. Overall output power, ~ 1.5 W; bandwidth at 0.1 dB below peak, ~ 40 Mc/s; group-delay variation over 20 Mc/s, without compensation, ~ 1 mμsec.

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[14]) Experiment done by W. J. Smulders.

extremely stable in operation. *Fig. 5* shows the block diagram of a typical local oscillator of this kind [15]. The first stage is a crystal oscillator (*Cr*) whose frequency is in the region of 25 Mc/s, and the required frequency, which may be, for example, in the region of 4000 Mc/s, is obtained by means of a series of frequency multipliers (*M*). The last stages use EC 157 tubes, the final multiplication being six-fold. A complete local oscillator designed on this principle for use in Philips microwave radio links can be seen in *fig. 6*. The output power is of the order of 200 mW, which is sufficient both for the receiving and transmitting ends of a relay station.

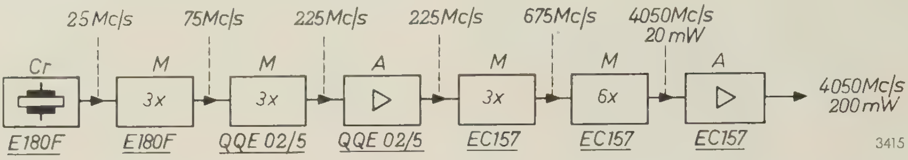

Fig. 5. Block diagram of a local oscillator [15] for microwave radio-link relay station. *Cr* crystal oscillator. *M* frequency multipliers, the last two fitted with EC 157 triodes. *A* amplifiers.

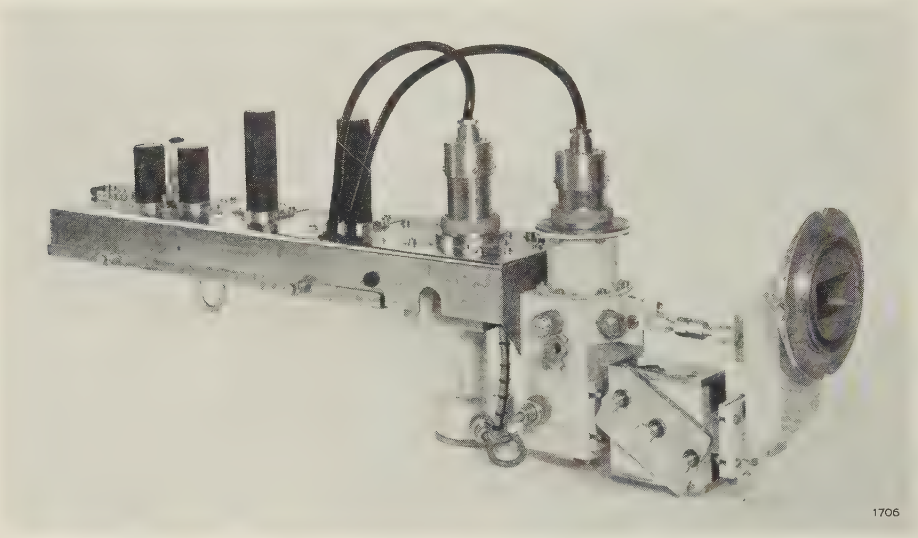

Fig. 6. A complete local oscillator, as in fig. 5, used in Philips microwave radio-link equipment.

The above-mentioned frequency of 4000 Mc/s is by no means an upper limit; the EC 157 can be used as a frequency multiplier at frequencies far higher. In an experimental set-up an EC 157 was driven at 4000 Mc/s by an input power of approximately 1 W. With the anode surrounded by a circuit tuned at the required frequency, all harmonics were obtained up to 24 000 Mc/s. One of the multiplier stages used is shown in *fig. 7*. The available power, particularly at the highest frequencies, is not large but is more than enough to be used, for example, in frequency standards. With the 5 cm, 1 W triode, excited at say 8000 Mc/s, it may be possible to reach the millimetre waveband.

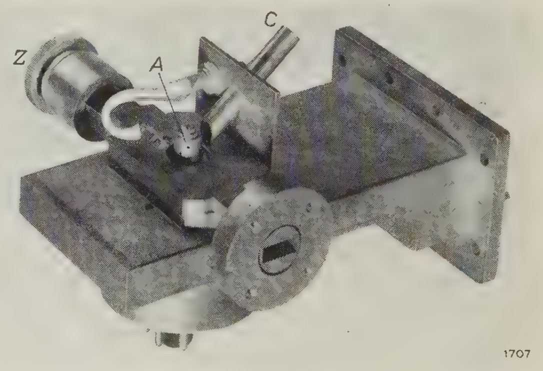

Fig. 7. Frequency multiplier stage equipped with an EC 157 triode, capable of producing 20 000 Mc/s by multiplying 4000 Mc/s. *A* anode. *C* nozzle for cooling-air supply. *Z* knob for adjusting tuning plunger. Driven by about 1 W at 4000 Mc/s the available power at the output is approx. 0.1 mW at 20 000 Mc/s.

Applications in oscillators

The local oscillator described above contains, as frequency multipliers, many electron tubes, which may be regarded as so many potential sources of

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[15] This design, like the mixer circuit in fig. 11, was developed by H. J. Kramer in the microwave radio-link laboratory of the N.V. Philips' Telecommunicatie-Industrie at Huizen.

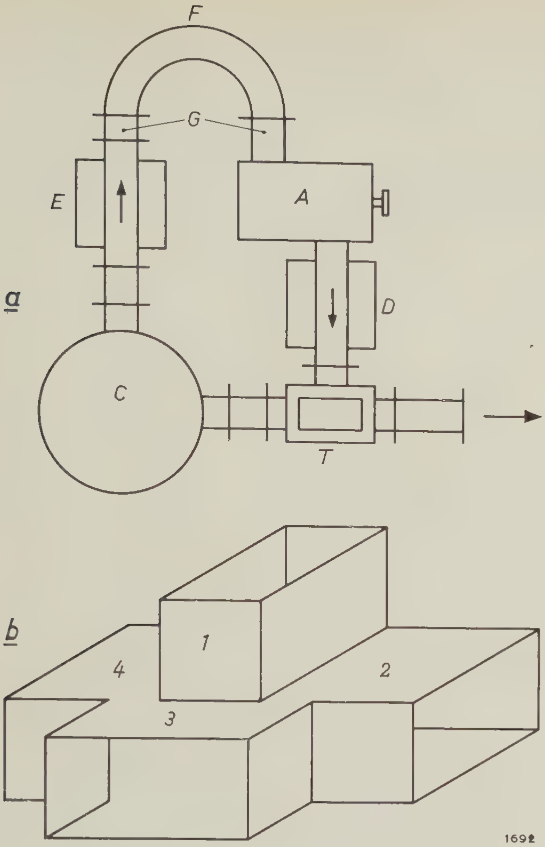
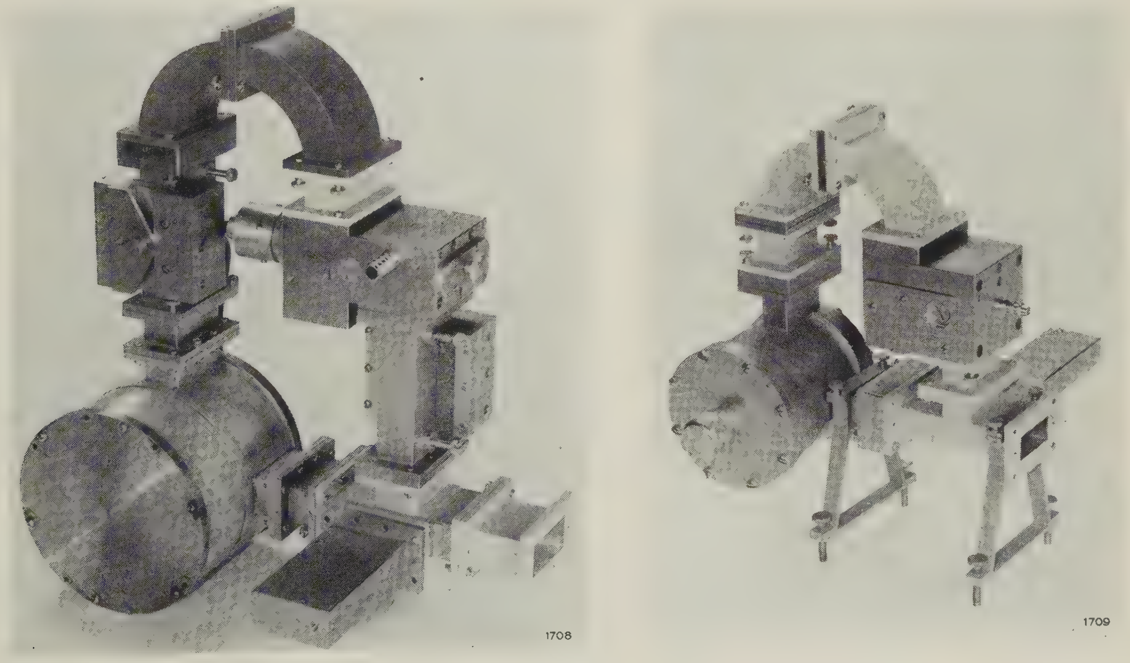


Fig. 8. *a*) Diagram of a 4000 Mc/s oscillator with an EC 157 triode. *A* amplifier. *C* cylindrical cavity resonator of "Invar", filled with dry air. *T* magic-tee (see *b*) as output coupler. *D*, *E* isolators. The phase of the signal returning via the curved section *F* to *A* is adjusted by means of spacer sections *G*. *b*) Magic-tee form of hybrid junction. Branches *1* and *3* have no corresponding dimensions parallel. In a magic-tee an electromagnetic wave entering one of the four branches *1*, *2*, *3* or *4* will divide equally between only *two* of the other three branches: paths *1*→*3*, *3*→*1*, *2*→*4* and *4*→*2* are not possible. (See for example G. C. Southworth, Principles and applications of waveguide transmission, Van Nostrand, New York 1950, pp. 339-340.)

breakdown. Attempts have therefore been made to design an oscillator which, with only one tube, will oscillate directly at the required frequency just as stably as a crystal oscillator. The components for such a local oscillator were available: the EC 157 triode can compete in every respect with tubes for lower frequencies, and with the aid of a cavity resonator the same high Q for high frequencies can be achieved as with the crystal. As the material for this resonator we can use "Invar" which, like quartz, possesses a very low thermal coefficient of expansion (both about 10^{-6} per °C). If the same care is paid to the filling of the resonator as to the envelope around the crystal, similar results may be expected.

Fig. 8a shows a diagram, and *fig. 9a* a photograph, of a 4000 Mc/s oscillator built on these lines. The amplifier *A* and the cylindrical cavity *C* are incorporated in a closed loop of waveguides. They are both terminated by an isolator (*D* and *E*, respectively). The circuit further contains a magic-tee *T* as the output coupler. In this junction the output power of the amplifier is split in two (see caption to fig. 8*b*): one half is fed to the output waveguide,



a *b*

Fig. 9. *a*) Oscillator for 4000 Mc/s fitted with an EC 157, built on the principle of fig. 8*a*. *b*) Similar oscillator, but for 6000 Mc/s and fitted with a 5 cm, 1 W triode. Both oscillators, *a* and *b*, are reproduced on the same scale.

the other half returns via the cavity to the input of the amplifier. A matched termination of the fourth branch of T absorbs any waves reflected from the cavity or the load.

Calculation has shown that the oscillator reaches optimum stability when the anode bandwidth of the amplifier is about 130 Mc/s and the attenuation of the resonant cavity 4.6 dB. The attenuation introduced by the output coupler amounts to 3 dB, and that due to each of the two ferrite isolators is about 1 dB. Since the loop gain must be unity, the operating point of the tube should be such that the gain is 9.6 dB; at this gain, and at the bandwidth of 130 Mc/s, its output power is 950 mW. The isolator D reduces this to 750 mW, half of which, i.e. 375 mW, is the available power at the output.

The Invar cavity has very smoothly finished interior walls. It is excited in the TE_{012} mode and is so designed as to give a loaded Q of 22 000. The cavity contains dry air, and is sealed by thin sheets of mica at the waveguide flanges and a synthetic-resin seal along the edges of the lids.

The oscillator is aligned as follows. First of all the cathode side of the amplifier is matched at the required frequency, and at the anode side the bandwidth is adjusted to about 130 Mc/s. The anode voltage is 200 V, the anode current 60 mA. After the complete oscillator loop has been assembled, the bent waveguide F (fig. 8*a*) is removed and a signal of the desired frequency is applied to the amplifier input. A detector is connected to the isolator E, and the amplifier and resonant cavity are then tuned exactly. Finally the curved section F is replaced and the phase of the returning signal is regulated by varying the loop-length by means of spacers G until the output power is maximum. If necessary the frequency can be finely adjusted with a trimming screw in the resonator.

The stability of the frequency during variation of the various parameters is measured by a beat-frequency method. Some results at 4000 Mc/s are given in *Table II*. The heater voltage proves to be the most critical of the parameters; the anode voltage has only a very minor influence. After exchanging the tube, only the anode side being retuned to maximum output power, it was found that the frequency had moved only 10 kc/s from the original value.

Fig. *9b* shows a similar oscillator for the frequency band around 6000 Mc/s, equipped with a 5 cm, 1 W type triode. Since the internal feedback in this tube is much lower than in the EC 157, isolators are not needed here.

Table II. Stability at 4000 Mc/s of the oscillator in fig. 9*a*.

Varying parameter	Variation of frequency with increasing parameter	
	Absolute	Relative
Temperature (25-65 °C)	+1.4 kc/s per °C	$+0.35\times10^{-6}$ per °C
Heater voltage	∼ −0.06 kc/s per mV	$\sim -1\times10^{-6}$ per %
Anode voltage	+0.05 kc/s per V	$+0.025\times10^{-6}$ per %
Anode current	+1 kc/s per mA	$+0.15\times10^{-6}$ per %

A triode can be used successfully as an oscillator not only at the design frequency and below, but also at far higher frequencies. Up to the design frequency the $\frac{1}{4}\lambda$ mode of the anode resonant cavity will be used; at much higher frequencies this is no longer possible because the resonant cavity would have to be smaller than is compatible with the tube dimensions. Nevertheless, the tube can be tuned to thees high frequencies by shifting the short-circuiting wall of the resonator half a wavelength outwards, thus using the $\frac{3}{4}\lambda$ mode. *Fig. 10* shows an experimental oscillator with an EC 157, which, on the principle described, is capable of generating frequencies up to about 8000 Mc/s.

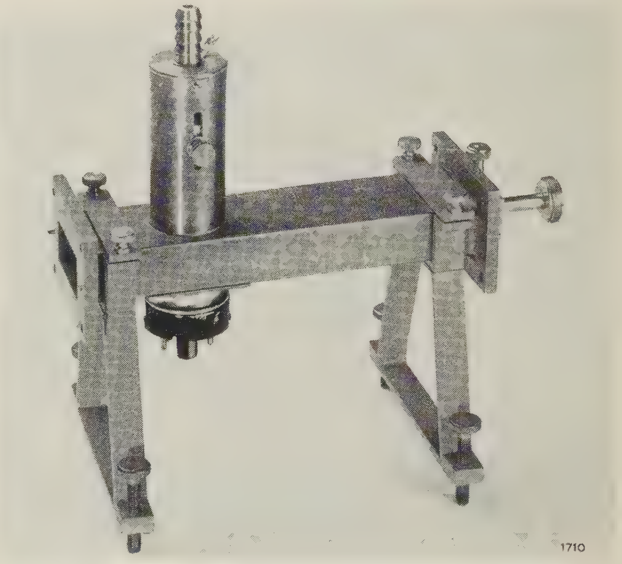


Fig. 10. Experimental oscillator with an EC 157 triode, with which frequencies up to about 8000 Mc/s are reached by using the $\frac{3}{4}\lambda$ mode of oscillation of the anode resonant cavity.

Mixing

If we apply to the input waveguide of an EC 157 amplifier the high-frequency signal from a local oscillator (LO signal) and apply to the cathode an IF signal, sum and difference frequencies appear at the anode. By tuning the anode circuit to one of these, e.g. the sum frequency, we obtain a mixer stage. A mixer of this kind is used in Philips microwave radio-link equipment for the purpose of transposing the signal from the intermediate frequency (70 Mc/s) to one of the radio-frequency bands at 4000 Mc/s. The circuit diagram is shown in *fig. 11* [15]), and the data are given in the caption. The anode circuit is followed by a three-cavity filter, which suppresses the LO signal and the unwanted sideband (difference frequency).

The anode current of the EC 157 is fairly small, being only 25 mA, at which value considerable AM-PM conversion (see p. 18) occurs. This is almost entirely compensated, however, by the AM-PM conversion of the driving tube. Since AM-PM conversion is mainly due to variation of the input capacitance with the magnitude of the driving signal, its sign in an earthed-grid arrangement will be opposite to that in an earthed-cathode arrangement. The AM-PM conversion of the driving tube is positive and that of the triode negative, and both are about 2°/dB, so that by suitable design of the circuit we can build a mixer stage capable of handling fairly strong signals and nevertheless virtually free of AM-PM conversion ($< 0.2°$/dB).

The triode as a mixer is also used in a straight-through relay station in microwave radio links for imparting to the signal the earlier-mentioned frequency shift of 213 Mc/s (see fig. 4). The RF signal is applied to the input of an EC 157 amplifier, and a sufficiently strong LO signal of the desired shift frequency is applied to the cathode (*fig. 12*). The amplitude-modulated RF signal is thus mixed with the shift frequency, which again gives rise to two sidebands. The sideband required is then selected by suitable filters.

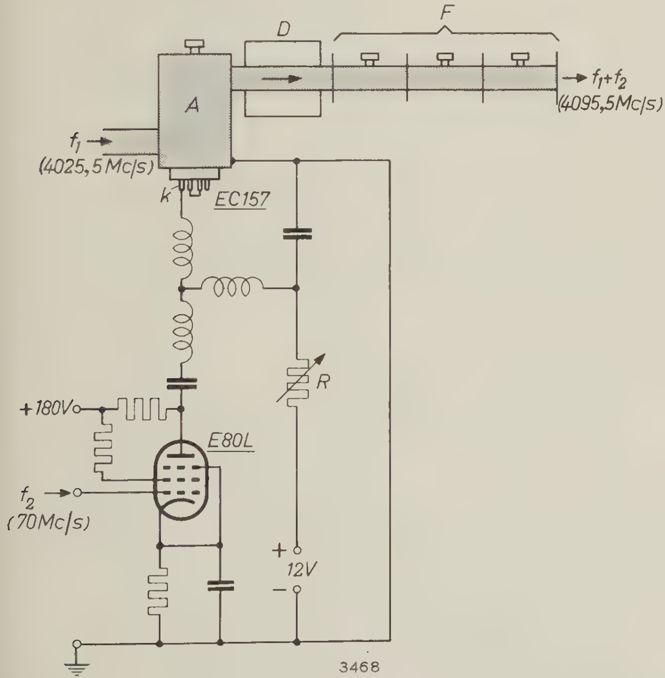


Fig. 11. Circuit [15]) for an EC 157 used as a mixer for transferring the modulation of an IF signal (f_2) to an RF signal (f_1). *A* amplifier. *D* isolator. *F* three-stage filter. *k* cathode lead of EC 157.

Conversion gain	3 dB
Bandwidth at 0.1 dB below peak	25 Mc/s
Input power of EC 157	200 mW
IF signal on grid of E 80 L (r.m.s.)	0.35 V
AM-PM conversion	<0.2 °/dB
Anode voltage of EC 157	180 V
Anode current of EC 157 (variable by means of *R*)	25 mA

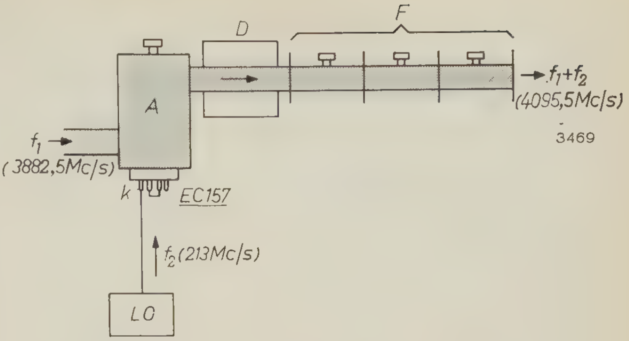


Fig. 12. Circuit for an EC 157 used as a mixer to produce a frequency shift of the RF signal in the straight-through relay station shown in fig. 3. *LO* local oscillator. Other symbols as in fig. 11.

Conversion gain	1 dB
Bandwidth at 0.1 dB below peak	50 Mc/s
LO power	500 mW
Anode voltage of EC 157	180 V
Anode current of EC 157	20 mA

Limiting

If a microwave triode is biased to operate at a low anode current and the input power increases, a situation is soon reached where the output power can rise no further. In this saturation region the tube may thus be expected to function as a limiter.

We can put this to the test by measuring the output power P_o as a function of the input power P_i; for the EC 157 this results in curve *A* in *fig. 13*. With increasing P_i this curve does not become flat, as we should wish for a limiter, but, at a larger driving signal, exhibits a very sharp dip. This effect is due to the passive feedback admittance between the cathode and anode circuits [16]). What in fact happens? The output power may be regarded as consisting of two parts. One part comes from the

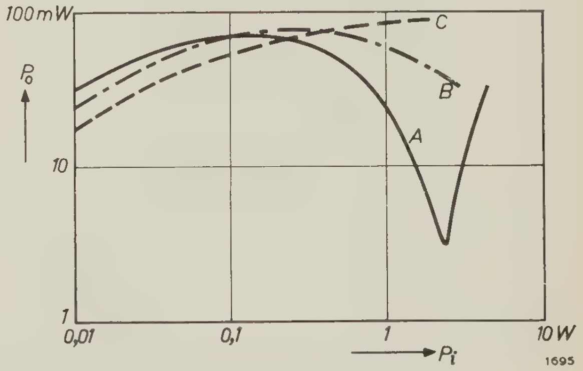


Fig. 13. The EC 157 as a limiter: output power P_o as a function of input power P_i. Anode voltage 200 V, anode current 10 mA. Curve *A*: amplifier not neutralized, curve *B*: partly neutralized, and curve *C*: completely neutralized.

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[16]) G. Diemer, Passive feedback admittance of disc-seal triodes, Philips Res. Repts. **5**, 423-434, 1950. See also the articles referred to under footnote [1]), p. 323, and [2]) pp. 152-154.

amplifying action of the triode. This part, which predominates in the case of small driving signals, approaches a constant value as the signal increases. The other part arises from the passive feedback admittance and has no saturation value, but continues to increase linearly with the input power. A point is reached, therefore, where these two parts are equal, and what happens then depends only on the phase relationship. Evidently in our case the two parts are virtually in antiphase, resulting in almost complete cancellation.

To avoid this undesirable effect we have tried to introduce between the anode and cathode cavities additional feedback, just sufficient to neutralize the existing feedback (internal neutralization). This is done, as shown in *fig. 14*, by passing a double coupling loop L of suitable dimensions through the partition between the two resonant cavities. In this way it proved possible to neutralize an amplifier completely. The result is curve C in fig. 13. It can be seen that the neutralization also results in a drop in gain for small P_i. The effect of partial neutralization will be to shift the dip towards a higher P_i. We then obtain curve B.

For use as a limiter the optimum curve lies somewhere between B and C. A drawback is that, with stronger neutralization, the flat region also shifts towards a higher P_i. In that region, then, the triode functions well as a limiter. The output power can remain constant within 0.4 dB for an input-power variation of 10 dB [17]).

Amplitude modulation

Even in microwave telecommunications, where amplitude modulation is not usual, it may be necessary in certain circuits to modulate a signal in amplitude, as for example in the circuit described in the next section. An amplitude modulator is also often indispensable in laboratory test equipment. Its operation is actually the same as that of the mixer discussed above, but in accordance with general usage we shall confine the term "modulation" to those cases where the modulating signal is an audio or video signal.

Depending on the strength of the RF signal, different circuits have to be used to minimize distortion. For weak signals it is sufficient to introduce the signal as a voltage source of low internal resistance in series with the cathode resistor, as shown in *fig. 15*. As long as the driving signal is not too strong, this voltage modulaton can be applied up to a modulation depth of 80%.

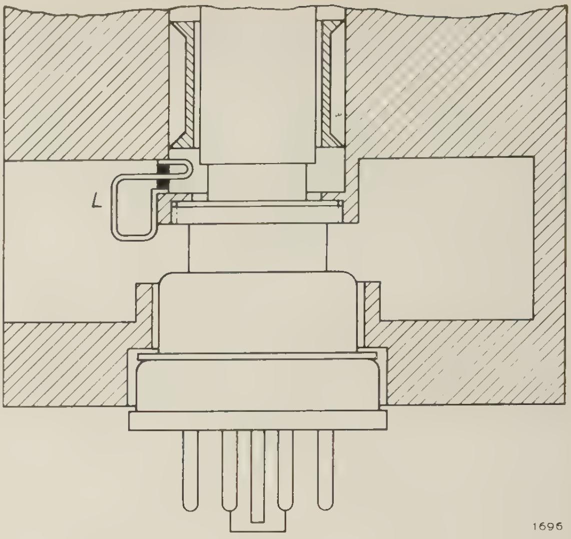

Fig. 14. Neutralization of an EC 157 amplifier by means of a double coupling loop L through the partition between anode and cathode resonant cavities.

Where a tube is operated close to the saturation level, however, current modulation is found to give the best results. For this purpose the signal source should be a current source of high internal resistance connected in series with the cathode resistor. The circuit adopted in practice is shown in *fig. 16*. The modulating voltage is applied between grid and cathode of a power pentode, type EL 86, connected in series with the amplifying tube. With this circuit an EC 157 almost driven to saturation can be modulated to a depth of about 90%.

In both cases the variation of the output *power*, and not of the output *voltage*, is proportional to the modulating voltage. This means that the detector must have a square-law characteristic.

Frequency modulation

Frequency-modulated microwaves are difficult to obtain with only one microwave triode used as an

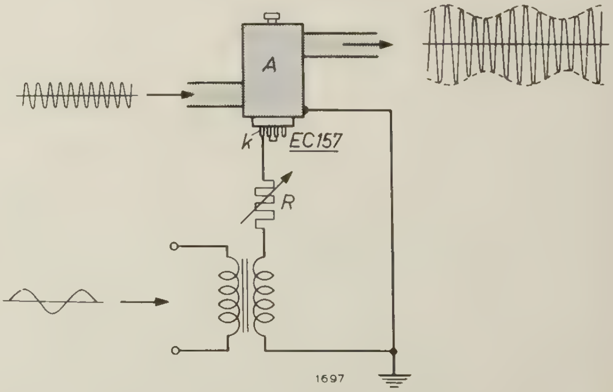

Fig. 15. Amplitude modulator with an EC 157 for modulating small signals. A amplifier. k cathode lead. $V_a = 200$ V, $I_a = 20$ mA (variable by means of R).

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[17] The AM-PM conversion of these and similar limiters has not yet been investigated.

oscillator. The oscillator frequency is then almost entirely determined by the tuned circuits, and although the frequency can be varied to some extent by varying the anode voltage and anode current, the circuit is not attractive for frequency modulation because the frequency deviation attainable is small and because of the large amplitude modulation which it involves.

The situation is different, however, if *two* tubes are used in a special circuit, the principle of which is illustrated in *fig. 17*. Two EC 157 amplifiers *A* and *B* are interconnected by waveguides (bold lines in the figure) via two magic tees T_1 and T_2. The path T_1AT_2 is made a quarter wavelength longer than the path T_1BT_2. Furthermore, T_1 and T_2 are directly interconnected via *PQ*.

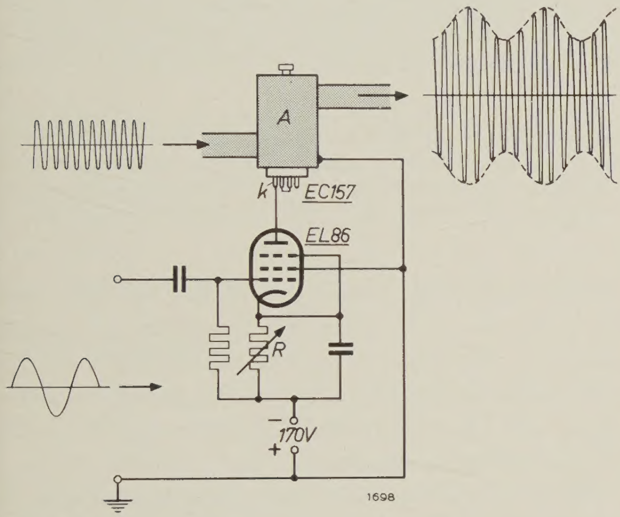

Fig. 16. Amplitude modulator with an EC 157 for modulating strong signals. Symbols as in fig. 15. $V_a = 200$, $I_a = 60$ mA.

To explain the operation of the circuit we shall imagine that the connection between T_1 and T_2 is broken for a moment at *P* and *Q*. An electromagnetic wave entering *Q* divides in T_1 into two waves of the same power and the same phase, one travelling to the left and the other to the right. These two waves are equally amplified in identical amplifiers *A* and *B*, and thus arrive in T_2 with equal amplitudes and, owing to the $\frac{1}{4}$-wavelength extra pathlength via *A*, with a phase difference of 90°. They can therefore be represented by the complex quantities α and β in *fig. 18*. In T_2 the waves α and β combine in such a way that a wave $\alpha + \beta$ travels towards *P* and a wave $\alpha - \beta$ along *S*.

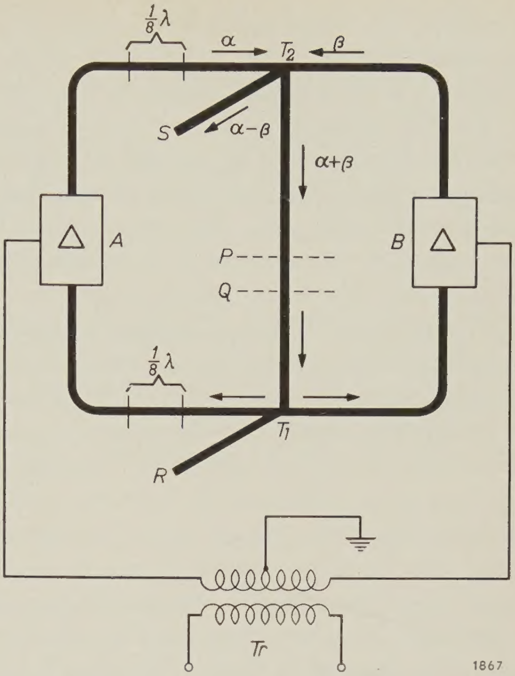

Fig. 17. Illustrating the principle of a frequency modulator with two EC 157 triodes. The thick lines represent waveguides. *A*, *B* amplifiers. T_1, T_2 magic-tees. The branch *R* has a matched termination, the branch *S* functions as output. *Tr* centre-tapped transformer (input).

Now suppose the waves α and β are not only amplified in *A* and *B* but also modulated in amplitude in push-pull, as represented diagrammatically by the centre-tapped transformer *Tr* in fig. 17. In practice this amplitude modulation can be effected by one of the methods described under the previous heading; the instantaneous output powers of the amplifiers then become varied in the ratio $(1 + m \sin pt) : (1 - m \sin pt)$, where m is the modulation depth and p the angular frequency of the modu-

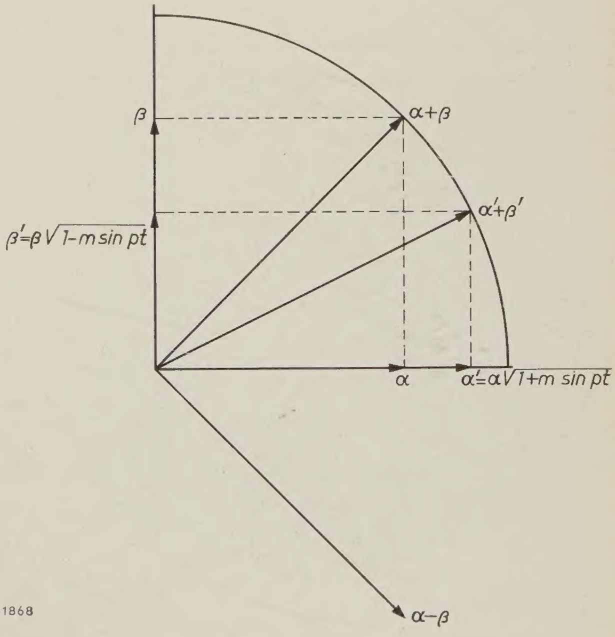

Fig. 18. Vector diagram illustrating the occurrence of phase modulation when the amplifiers *A* and *B* in fig. 17 are push-pull amplitude-modulated.

lating signal. The amplitudes of the waves arriving in T_2 are thus given by $\alpha' = \alpha\sqrt{1 + m \sin pt}$ and $\beta' = \beta\sqrt{1 - m \sin pt}$, and those at the output of T_2 by $\alpha' + \beta'$ and $\alpha' - \beta'$.

It is easy to show that the moduli of both these output waves are equal and constant, and that the arguments vary with time. The output waves from T_2 are therefore not modulated in amplitude but only in phase: the vector points of $\alpha' + \beta'$ and $\alpha' - \beta'$ thus move, during a modulation period of the amplifiers, along arcs of the same circle in fig. 18.

If we now connect P again with Q, the circuit will start to oscillate, and it will do so at a frequency such that the total phase shift in the feedback loop is exactly a whole multiple of 2π. Since the modulating voltage causes the "phase length" of the circuit to vary, the frequency at any given moment will be such as to satisfy the just-mentioned phase condition. In this way, then, we have obtained a frequency-modulated oscillator. The branch S of T_2 functions as the output, and the branch R of T_1 has a matched termination.

To produce a large frequency deviation it is necessary that the phase variation introduced should constitute as large a part as possible of the total phase shift in the loop. This amounts to keeping the whole feedback path as short as possible. A compact assembly is achieved by combining the two magic-tees and their connection PQ in a single block (*fig. 19*). The two amplifiers are normal amplifier units, except that input and output are at the same side. The amplifiers are fitted to the left and right of the block in fig. 19 (see *fig. 20*), and the difference in path length is obtained by means of spacer sections. The bottom opening at the front of the central block (R in fig. 17) has a matched termination, and the upper opening (S in fig. 17) is the output of the modulator.

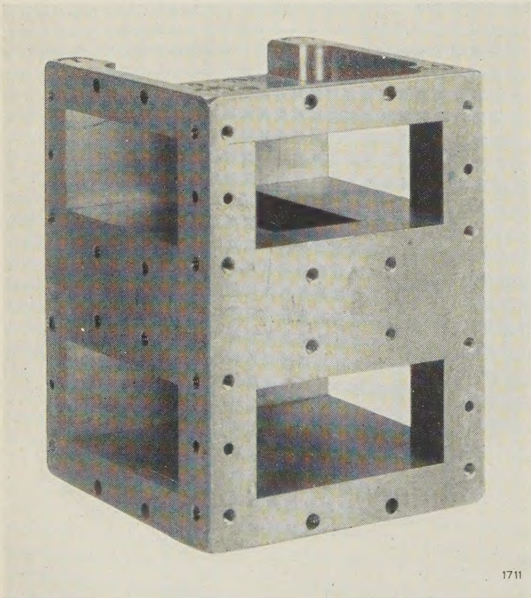

Fig. 19. Magic-tees T_1 and T_2 and their connecting waveguide (see fig. 17) combined in a single block, giving a compact construction of the frequency modulator.

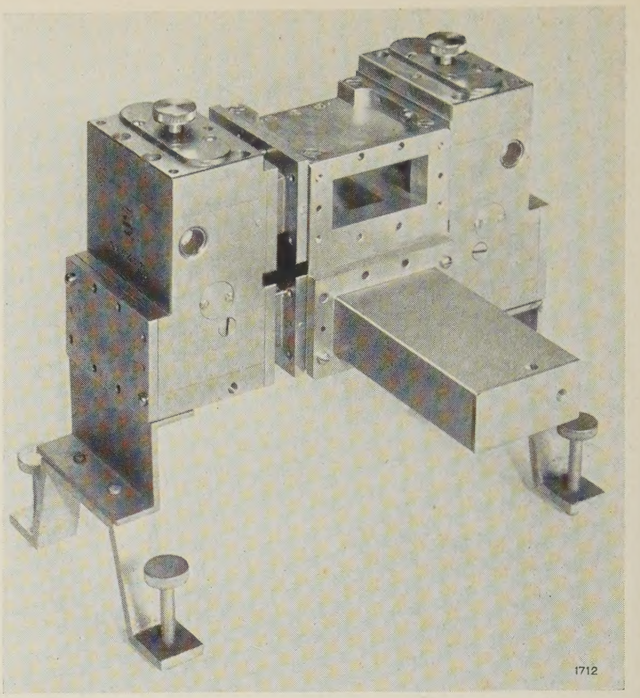

Fig. 20. Complete frequency modulator, built on the principle of fig. 17, using the block shown in fig. 19. The feedback paths of the two triodes are unequal in length and in part common. When the waves in the triodes are amplitude-modulated in push-pull, the output signal will be frequency-modulated.

To help shorten the feedback path, no isolators are fitted at the anode side of the amplifiers. This makes it necessary to neutralize the amplifier blocks, which is done by the method described above (fig. 14).

As a result of this special circuit arrangement, the output power of the frequency modulator is fairly high. As we have seen, the moduli of the output waves of T_2 are identical, so that the sum of the output powers of the amplifiers divides into two equal parts in T_2. Of these, one part is again split in T_1 into two equal parts, and the other is immediately available as output power. We may therefore expect the output power to be equal to that of a single tube driven by a signal strong enough to produce a gain of 3 dB. What the actual power will be depends on the setting of the modulator. The above reasoning was based on a phase difference of 90° between α and β. Plainly, by varying this angle we can play off the output power against the frequency deviation. If we make the angle larger we immediately increase the deviation, but from fig. 18 we see that

$\alpha + \beta$ then becomes smaller, that is to say, less power is fed back and the tubes will oscillate at a lower level. Conversely, a smaller angle gives a smaller deviation, but $\alpha + \beta$ is then larger, more power is fed back and the amplifiers oscillate more strongly.

We can make this process more effective if, at the same time, we vary the anode-circuit bandwidth of the amplifiers, depending on whether we want a large deviation or a large output power. *Table III* gives some results of measurements made at two settings representing roughly the extremes of deviation and output power.

Table III. Data on the frequency modulator in fig. 20, equipped with two EC 157 triodes, in two different settings.

	For large frequency deviation	For high output power
Anode voltage	200 V	200 V
Anode current	60 mA	60 mA
Anode-circuit bandwidth	150 Mc/s	60 Mc/s
Centre frequency	4000 Mc/s	4000 Mc/s
Maximum frequency deviation	60 Mc/s	15 Mc/s
Linear frequency deviation	20 Mc/s	5 Mc/s
Output power	0.6 W	3.6 W

High-level detection

In millimetre radar, where pulses of extremely short duration are used (about 5 mμsec), a very large bandwidth (about 200 Mc/s) is required in the IF section and video amplifiers in the receiver [18]). A bandwidth as large as this is very difficult, if not impossible, to achieve with conventional tubes. As far as the IF amplifier is concerned, the difficulty is solved by using a travelling-wave tube, operating at a frequency of 4000 Mc/s. A normal crystal detector cannot, however, handle anything like the maximum output power of the travelling-wave tube without becoming overloaded. Considerable video amplification is therefore needed behind the detector in order to drive the cathode-ray tube.

This video amplification, which is very difficult to realize for such a large bandwidth, can be dispensed with entirely if the detection is done by an EC 157 triode, used as an anode-bend detector. The travelling-wave amplifier can boost the IF signal to about 1 W, and the triode detector, driven by this signal, delivers over the whole bandwidth video signals of about 10 V, which can be used for driving the cathode-ray tube directly. This arrangement is shown schematically in *fig. 21*.

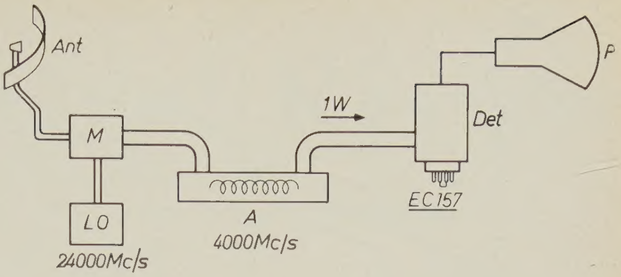

Fig. 21. The EC 157 used as an anode-bend detector for strong signals in millimetre radar. The IF amplifier *A* is a travelling-wave tube, which supplies a signal of about 1 W at 4000 Mc/s to the triode detector *Det*. The detector output signal is strong enough, at the required bandwidth, to drive the cathode-ray tube *P* directly, i.e. without video amplification. *Ant* antenna. *M* mixer. *LO* local oscillator (24 000 Mc/s).

An ordinary amplifier block can be used for such a detector; the top cover is removed and also the inner conductor and plunger, in order to minimize stray capacitances. The circuit used is shown in *fig. 22*. The tube is biased by a fixed positive

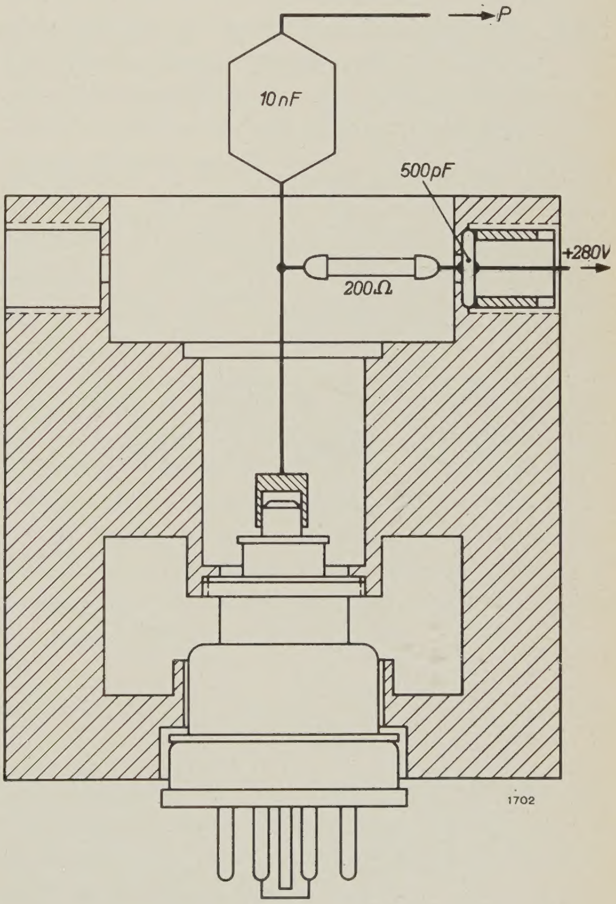

Fig. 22. Arrangement of EC 157 as anode-bend detector at 4000 Mc/s in a slightly modified amplifier block. The tube is biased to cut-off by a fixed voltage of +7 V on the cathode. The cathode-ray tube is connected at *P* (see fig. 21).

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[18]) See R. J. Heijboer, Millimeterradar, Ingenieur 71, E. 41-E. 48, 1959 (No. 14) (in Dutch).

potential of 7 V on the cathode, which corresponds roughly to the cut-off point. *Fig. 23* shows the video voltage on the anode as a function of the IF power at the input.

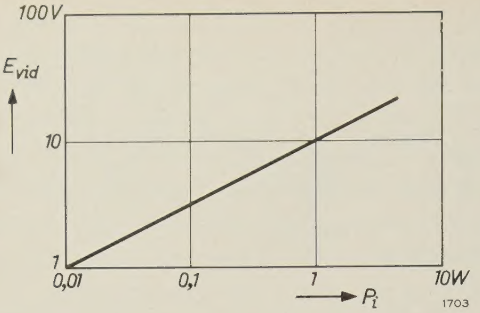

Fig. 23. Detection characteristic of the EC 157 used as an anode-bend detector in the arrangement shown in fig. 21: the video voltage E_{vid} on the anode is plotted as a function of the IF power P_{i} at the input.

Summary. Because of such features as simplicity, favourable phase characteristics and large product of gain and bandwidth, disc-seal triodes for centimetre waves are suitable and attractive for many and various applications, especially in microwave radio links. Tabulated details are given of the microwave triodes developed or in development at Philips, and their uses are discussed as amplifiers, frequency multipliers, local oscillators of extremely stable frequency (4000 and 6000 Mc/s), mixers, limiters, amplitude and frequency modulators and high-level detectors for millimetre radar.

ABSTRACTS OF RECENT SCIENTIFIC PUBLICATIONS BY THE STAFF OF N.V. PHILIPS' GLOEILAMPENFABRIEKEN

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2736: W. Kwestroo and H. A. M. Paping: The systems BaO-SrO-TiO_2, BaO-CaO-TiO_2 and SrO-CaO-TiO_2 (J. Amer. Ceram. Soc. **42**, 292-299, 1959, No. 6).

The solid phases formed at 1400 °C in air in the three-component systems BaO-SrO-TiO_2, BaO-CaO-TiO_2 and SrO-CaO-TiO_2 are described. Besides solid solutions of components with known structures, some new ternary compounds have been studied. The dielectric constants and loss factors of a number of specimens are given. Crystallographic data of the compounds $BaCaTiO_4$, $Ba_3Ca_2Ti_2O_9$ and $Ca_3Ti_2O_7$ and of the solid solution series $(Ba,Sr)_2TiO_4$ are presented. The preparation of the new compounds is described in detail.

2737: K. van Duuren, A. J. M. Jaspers and J. Hermsen: G-M counters (Nucleonics **17**, No. 6, 86-94, 1959).

The article describes some modern designs of Geiger-Müller tubes. Diversity of design provides G-M tubes for many uses: hollow anodes for 4π and liquid-flow counting, compact coincidence units and integrator tubes for inexpensive, reliable monitors and alarm devices.